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EDITORIAL

Where was the Telephone Invented?

FOR over seventy years there has been discussion as to who invented the telephone and even within the last month we have received a long letter from one of his descendants claiming that the honour really belongs to Alexander Bain, well-known as the inventor of the electric clock and of the chemical telegraph. This claim seems to be based on some family tradition and is probably due to the fact that the words telegraph and telephone were used in rather a confused way in the early days. This is brought out very clearly in Kelvin's report on the Philadelphia Centennial Exhibition, where he heard "to be, or not to be" on the electric *telegraph* and referred to an electric *telephone* sending four messages simultaneously on the Morse code.

Very few people now have any doubts as to the justification for regarding Alexander Graham Bell as the inventor of the telephone, but we have, within the last few weeks, received two letters from Canada suggesting that in the recent celebrations of the centenary of Bell's birth, too much credit was given to Boston and too little to Brantford as the birthplace of the telephone. It was to Brantford in Ontario that Bell's father and mother went from London in 1870, hoping that the climate would be beneficial to the health of Alexander Graham, whose two brothers had recently died of tuberculosis and whose own health was causing grave anxiety. There on Tutela Heights the family settled and there Bell's father and mother lived until 1881 when they followed their—by that time, celebrated—son to Washington. At Brantford Bell's health rapidly improved and he took a post in Boston, but he spent his holidays at Brantford and discussed his ideas with his father.

The experimental work and realization of the invention was carried out in Boston, but then when he returned to Brantford he carried out transmission experiments between Brantford and neighbouring towns. In 1916 a tablet was unveiled at 109, Court Street, Boston, with the inscription "Here the telephone was born June 2, 1875." In 1917 a monument was unveiled at Brantford "To commemorate the invention of the telephone by Alexander Graham Bell, in Brantford in 1874." Bell was present at the ceremony and naturally did his best to justify the claims of Brantford. He said "Brantford became my thinking place; here the telephone was invented; the first draft of the patent specification prepared; the proper relation of the parts of a telephone to enable it to be used. on long lines, worked out; and the first transmission of the human voice over miles of telegraph line actually accomplished. Here also the first public demonstration of this result was given to the world. I think, therefore, that Brantford is fully justified in considering itself as integrally associated

with the development of the telephone."

The long distance transmission referred to was on August 10, 1876, when by arrangement with the Dominion Telegraph Company he established one-way communication between Brantford and a neighbouring town of Paris, really only 8 miles away, via Toronto, a distance of 136 miles; this was certainly the first long distance telephone call in the world, but it was only one way and the reply was by telegraph. The first two-way long distance transmission was in the United States. On another occasion Bell said "It so happens that the telephone was invented in Canada at Tutela Heights, during my visit to my father and mother in 1874; but the first telephone was made in Boston in 1875, and all the early experiments were made there up to the time of the Centennial Exposition at Philadelphia. Thus the telephone was conceived in Brantford, reduced to practice in Boston, and became known to the world at the Centennial Exposition in Philadelphia."

Bell's father kept a diary, and during the Christmas holidays in 1874 there are such entries as: "Dec. 27. Long talk on multiple telegraphy and speech transmission. Al. sanguine." "Dec. 29. Talking half the night—motor and telephone." It was in Boston in the following June that, during experiments with the multiple telegraph, the slight mishap occurred that led to the construction overnight of Bell's first telephone. Where the telephone was invented depends entirely on the meaning attached to the word "invention." Long talks and discussions of schemes, however enthusiastic, can hardly be regarded as invention, and there was no suggestion of a patent specification until after the incident at Boston, for we have Bell's own statement that he drew up the first draft of the telephone patent specification at Brantford in September or October of 1875.

G.W.O.H.

CONCENTRIC LINE*

Critical Wavelength of the Higher Modes

By H. Bondi and S. Kuhn

IN normal practice, the field in a concentric-line feeder is entirely in a plane transverse to the axis. With the use of very short waves there arises the possibility of other modes in a concentric line. When this happens there will, in general, be a complicated system of standing waves in the line, and in many applications this is undesirable. The purpose of this note† is to give curves showing, for any dimensions of the line, the critical wavelengths at which these modes occur.

The higher modes are similar to those in a circular waveguide and fall into two classes, the E-modes and the H-modes. In an E-mode the electric field has a longitudinal component, but the magnetic field is wholly transverse; in an H-mode the magnetic field has a longitudinal component, but the electric field is wholly transverse. They are further classified according to half the number of nodes encountered in the field in a full revolution round the axis and the number of nodes encountered in the longitudinal or tangential electric field in going radially from one conductor to the other (actually one plus the latter). Thus an H_{11} mode has two nodes in going round a full revolution and no nodes of tangential electric field in going radially from one conductor to the other.

We consider a concentric line filled with a uniform dielectric of dielectric constant ϵ and unit permeability, referred to vacuum. In the free dielectric waves of frequency f are propagated with a wavelength λ_e given by

$$\lambda_e = c/f\sqrt{\epsilon} = \lambda_0/\sqrt{\epsilon},$$

where c is the velocity of light in vacuo, viz, 3×10^{10} cm/sec, and λ_0 is the wavelength in vacuo. While the wavelength of the normal mode of propagation in the concentric line is also λ_e, the wavelength λ_g of any higher mode is greater than this, and is given by the formula

$$\frac{1}{\lambda_g^2} = \frac{1}{\lambda_e^2} - \frac{1}{\lambda_{cr}^2}$$

where λ_{cr} is a length characteristic of the

* MS. accepted by the Editor, June 1946.

†Based on Admiralty Signal Establishment Report M.423 of May 1942.

concentric line, which depends only on its dimensions and not on the nature of the dielectric. The critical frequency f_{cr} below which no propagation in this mode is possible is related to λ_{cr} by the formula

$$f_{cr} = c/(\lambda_{cr}\sqrt{\epsilon}) ;$$

in other words, λ_{cr} is the wavelength in the free dielectric of waves of the critical frequency, whereas in free space their wavelength is $\lambda_{cr}\sqrt{\epsilon}$. It will be seen that, compared with a vacuum in the concentric line, the presence of dielectric decreases the critical frequency by a factor $1/\sqrt{\epsilon}$.

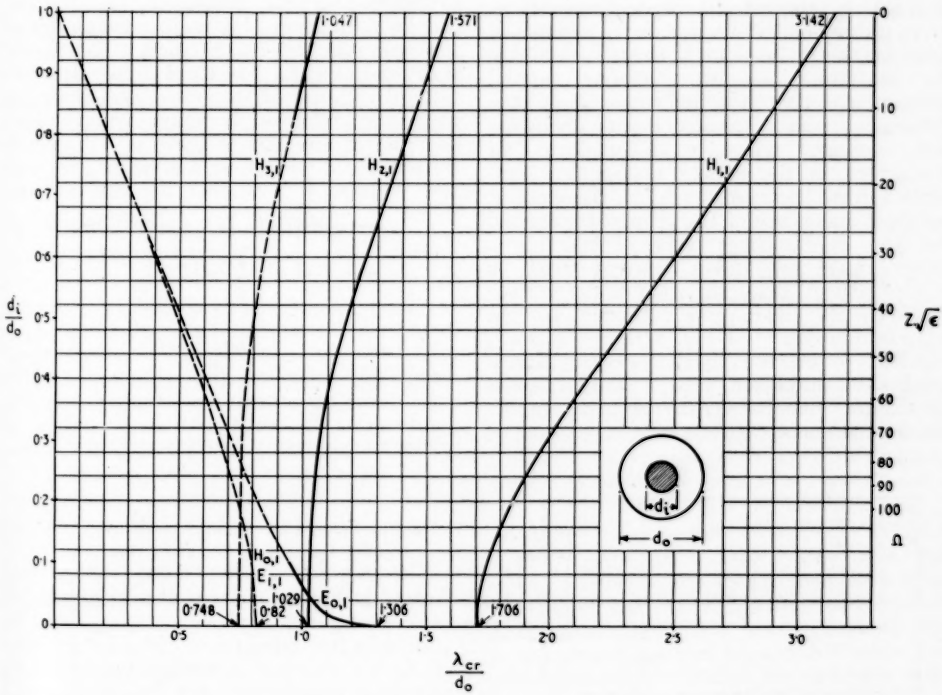
The normal mode does not fall into this classification, but for convenience it can be referred to as either the E_{00} or the H_{00} mode. It can exist at all wavelengths and its critical wavelength can be regarded as infinite.

The accompanying curves show, for the first few modes, the ratio λ_{cr}/d_0 of the critical wavelength (in free dielectric) to the outer diameter, plotted against the ratio d_i/d_0 of the diameters (or, equivalently against $Z\sqrt{\epsilon}$, Z being the characteristic impedance of the line for its normal mode).

They can also be used to determine the critical diameter for a given wavelength λ_e in the free dielectric and a given characteristic impedance, by interpreting λ_{cr}/d_0 as λ_e/d_{0cr}.

As the wavelength is diminished, the first new mode to appear in any concentric line is the H_{11} mode.

It will be seen that, to a very good approximation, the critical wavelength of this mode is equal to the mean circumference of the inner and outer conductors. This rule is deviated from by less than 3 per cent if the inner diameter is not less than a twelfth of the outer diameter, and the maximum deviation is 8·8 per cent.



Critical wavelengths of higher modes in a concentric line.

The work described in this paper was carried out in the Admiralty Signal Establishment and the paper is published with the approval of the Board of Admiralty.

STABILIZING DIRECT-VOLTAGE SUPPLIES*

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SUMMARY.—A graphical method of assessing the performance of the gas-discharge tube stabilizer circuit is described, and a means of ensuring that the tube "strikes" with low supply voltages is indicated.

1. Introduction

THE h.t. voltage for electronic equipment is usually obtained by using a valve rectifier supplied from a.c. mains, or directly from d.c. mains. In both cases the voltage is subject to random variations from its nominal value because of changes of load on the mains, and it is often desirable to stabilize the h.t. voltage supply, particularly that for instruments of the laboratory type. Various circuits giving an h.t. voltage relatively independent of variations in the supply voltage are described in the literature [1, 2, 3, 4, 5]; the simplest of them makes use of the property possessed by the gas-discharge tube of carrying a large range of direct current with a comparatively small variation of the voltage between its anode and cathode. In the standard circuit [1,2] of Fig. 1 the voltage V_2 applied to the load is that existing across such a "stabilizer tube," and a series resistance r between it and the supply limits the current I_3 in the tube to a value

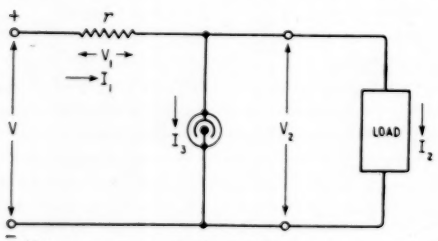

Fig. 1. *The basic stabilizer circuit.*

below its safe maximum. In this paper a graphical method of assessing, for various values of r, the degree of stabilization obtained with respect to input voltage variations and changes of load current is discussed; a simple method of ensuring that the stabilizer tube "strikes" when the supply voltage V is below the striking voltage is also described.

*MS. accepted by the Editor, October 1946.

2. The General Problem

In order to deal with the stabilizing properties of the circuit of Fig. 1, the graphical solution of a more general circuit problem will first be obtained. The circuit is shown in Fig. 2. The three circuit elements (a), (b), (c), are all supposed non-linear, so that the relation between voltage and current for any one of them is represented by a curve, and not by a straight line, and the voltage V available depends on the

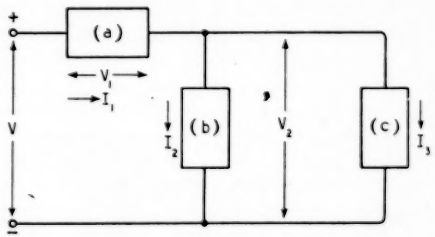

Fig. 2. *Generalized voltage stabilizer.*

current I_1. The values assumed by V, V_1, V_2, I_1, I_2, and I_3 are required.

The graphical solution of the problem is indicated in Fig. 3. Curve I shows the relation between V_2 and I_3; curve II, obtained by adding to the ordinates of curve I the values of I_2 for the corresponding values of V_2 shows the relation between V_2 and $(I_2 + I_3) = I_1$. Curve III represents the variation of V with I_1, and IV, obtained by subtracting from V the values of V_1 for various values of I_1, represents the variation of $(V-V_1)$ with I_1. Curves II and IV intersect at A where $(V - V_1) = V_2$. Its ordinate gives the current I_1, made up of AB (I_2) and BC (I_3), and DE gives the value of V, made up of DA (V_2) and AE (V_1).

If now the V—I_1 curve takes up a new position because of a change in supply

voltage, the effect on the position of A, and hence on the distribution of voltage and current in the circuit, is readily assessed. It is also clear that disconnection of the element drawing the current I_2 brings the point A to F, the point of intersection of curves I and IV.

Though the problem does not arise in the present considerations, it is worth noting that the method is easily extended to the cases in which there are further elements in parallel with either (b) and (c), or with (a). In the first of these, the I_1—V_2 curve is obtained by extending the above procedure; in the second, the curve relating V_1 to I_1 is first obtained in a similar way, the value of V_1 is found as before, and the currents in the elements in parallel with (a) are obtained from the curves used to construct the $I_1 - V_1$ curve. The method of solution here outlined appears to have some advantage over that given by Harnwell[6] for solving a similar type of problem in which only one of the elements is non-linear.

3. Circuit Design and Action

3.1 *Data required.*

In the circuit of Fig. 1, the elements (a), (b), (c) of Fig. 2 are replaced by the resistance r, the load and the stabilizer tube, respectively. The method of Section 2 will be applied to find the value of r appropriate to a given load and the supply voltage available, and to assess the degree of stabilization achieved by the circuit under

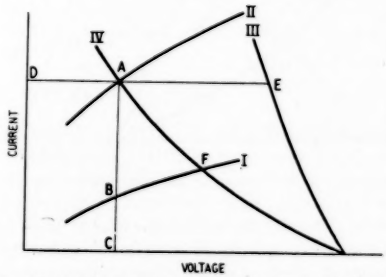

Fig. 3. *Graphical solution of the general three-element circuit.*

different conditions. Two types of direct voltage supply will be considered. In the first of these the d.c. mains are used, and it is assumed that the supply voltage V is independent of the current I_1 drawn from it. In the second type the supply is derived from a rectifier system, and the variation of V with I_1 is taken into account. The following characteristics will be supposed available, either from previous experiment or from manufacturers' data:

(a) *Stabilizer Tube:* variation of voltage V_2 with current I_3, maximum safe current, minimum current required to maintain the struck condition, and striking voltage.

(b) *Supply:* nominal voltage, and extreme departures from the nominal value; variation of direct voltage V with current I_1 when derived from a rectifier.

(c) *Load:* current drawn at the stabilized

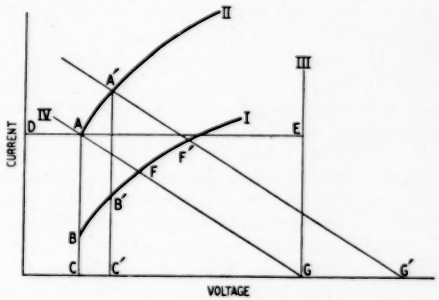

Fig. 4. *Determination of conditions for a d.c. mains input supply.*

voltage V_2, its small variation with the (small) variations in V_2 arising from changes in the supply voltage V, and the limits between which the current may be expected to vary owing to the addition to, or removal from, the system of items of equipment.

3.2 *Graphical procedure (d.c. mains supply).*

The highest permissible value of r, the series resistor, is that which, with the supply voltage at its lowest and the load current at its maximum, just avoids making the stabilizer voltage and current fall below the minimum values needed to maintain conduction. Fig. 4 shows the procedure for determining this value of r. It will be shown below that

(a) a lower value of r, with the same load current, will give worse stabilization,

(b) if the load current is decreased below the maximum, the behaviour of the system using the chosen value of r is easily assessed, and that the degree of stabilization then obtained differs little from that available when the maximum current is being taken.

Thus the highest permissible value of r

gives best operation, and in the subsequent discussion it will be assumed that this value is being used.

As in Fig. 3, curve I represents the stabilizer characteristic (I_3—V_2); curve II is obtained by adding to curve I the currents I_2 corresponding to the various voltages V_2 across the load. The ordinate through B, the point corresponding to minimum stabilizer voltage and current, meets curve II in A, the ordinate of which gives the total current I_1 for the voltage corresponding to A. If G represents the lowest supply voltage, the vertical line through G (III) corresponds to curve III of Fig. 3, and the straight line GA corresponds to curve IV. The horizontal line AE through A, meeting curve III in E, thus gives the voltage drop required in r at the current I_1 corresponding to A, and r is given by AE/EG, the reciprocal of the slope of GA.

If G′ corresponds to the highest supply voltage, the relation between V_2 and I_1 at this voltage is now known, since r has been found; it is represented by the line through G′ parallel to GA, meeting curve II in A′. Then the ordinate A′C′, meeting curve I in B′, gives the total current I_1 now flowing, made up of B′C′ (I_3) and B′A′ (I_2), and the abscissa at C′ gives the new stabilized voltage. A similar procedure gives I_2 and I_3 at the nominal supply voltage.

3.3 *Further deductions.*

The diagram also illustrates certain other points:

(a) When the maximum value of r has been chosen for a given maximum load current, the behaviour of the stabilizer voltage for a smaller load current is indicated by the point of intersection with GA and G′A′ of a line drawn in the same way as curve II. Since this line will usually lie between curves I and II, and will have a slope rather less than that of I, it follows that

(i) the limits between which V_2 varies are higher in absolute value than before,

(ii) the limits are somewhat further apart than before. Since curves I and II are almost parallel in practice, this has little effect on the degree of stabilization obtained.

(b) If the load is removed altogether, making I_2 zero, the new limits of stabilized voltage and stabilizer current are given by the points F and F′ where GA and G′A′ cut curve I. Now the highest stabilizer current occurs when there is maximum supply voltage and zero load current. It is thus immediately evident whether this exceeds the maximum safe current for the stabilizer; methods of reducing this current are considered in Section 5.2 below.

(c) If a value of r below the maximum is chosen, then

(i) GA and G′A′ will have a greater slope, and the horizontal separation of A and A′ will be further increased, so that the degree of stabilization obtained will not be as good as with maximum r,

(ii) the current limits for the stabilizer tube are both higher than before, and overloading occurs more readily.

(d) The greater the difference between the minimum stabilized voltage and the minimum supply voltage, the smaller is the variation of stabilized voltage with a given supply fluctuation GG′, since GA and G′A′ will then be more nearly horizontal and the horizontal separation between A and A′ for a given value of GG′ will decrease.

(e) The smaller the maximum load current for which the circuit is designed, the better is the stabilization achieved; for under these conditions the line II has a smaller vertical displacement from I, GA and G′A′ are more nearly horizontal, and the point A′ is again nearer A.

(d) and (e) both lead to the conclusion

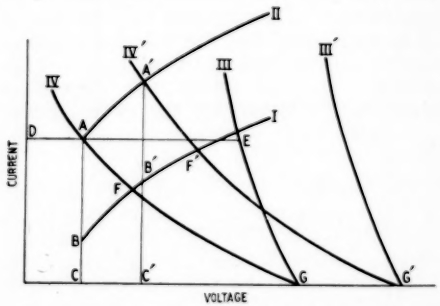

Fig. 5. *Conditions for an a.c. mains supply.*

that stabilization is best when the load current is small and the nominal supply voltage is high. Under these circumstances the power consumed in the load is small compared with that dissipated in the stabilizer tube and the resistance r, so that stability of voltage is obtained at the expense of economy in consumption.

3.4 *A.C. mains supply.*

The procedure of 3.2 for determining r now requires some slight modification, since

the direct voltage V available at the terminals of the rectifier system varies considerably with the current I_1 drawn from it. Fig. 4 is altered as follows:

Curve III becomes the I_1—V curve for the rectifier circuit for the minimum a.c. voltage input; the point E is found, and the value of r is deduced, as before. The line GA becomes the curve obtained by plotting $(V - rI_1)$ against I_1, and G′A′ is replaced in the same way by using III′, the I_1—V curve for maximum a.c. voltage input. Fig. 5 illustrates the resulting diagram, from which the degree of stabilization and the stabilizer current are obtained as in Section 3.2. The remarks made in 3.3 apply equally to this case.

3.5 *Non-linear element in place of r.*

It is possible to use a non-linear element in place of r if its current-voltage characteristic is such that it makes curve IV of Fig. 3 pass through the points G and A. The effect on the stabilizing properties of the circuit depends on the shape of the characteristic, as shown in Figs. 6 (a) and (b). If it is convex upwards (Fig. 6 (a), tungsten filament lamp[7]) then variations in the supply voltage produce smaller changes in the stabilized voltage than they would if the appropriate resistor were used. On the other hand, variations due to diminution of load are larger than with the resistor. If the curve is concave upwards (Fig. 6 (b), thyrite and similar materials[7]) these results are interchanged, and stabilization against change of load is better, and against change of supply voltage is worse, than when a resistor is used.

4. Stabilizer Striking

In order to obtain the stabilizing action it is essential that after switching on, the voltage applied to the stabilizer tube shall momentarily exceed the striking voltage so that conduction is ensured. It is usually necessary to modify the circuit of Fig. 1 to achieve this; the changes required are next considered.

If the character of the load is such that full load current is taken as soon as its H.T supply is switched on, it is unlikely that the load voltage will at any time exceed the striking voltage. Since r, V, and the I_1—V characteristic are known, a graphical determination of the resulting load voltage is easily carried out by the method of Section 3.3 (b). If it does not, a switch can be incorporated into the circuit so that the load can be added after the mains voltage has been applied to strike the stabilizer tube; the initial stabilizer overload which may occur need only be momentary. Where the load current is mainly that taken by indirectly-heated valves, the necessary action is obtained by switching on their heaters at the same time as the mains supply.

When the nominal d.c. mains voltage is below the striking voltage the use of a switch is clearly inadequate. To produce a voltage which will cause striking, the resistance r is replaced by an inductance L, with its resistance made up to r with a suitable series resistor, and a capacitor C is connected in parallel with the stabilizer tube and the load. It is shown in the Appendix that when the d.c. supply is switched on, the voltage across C surges above the mains voltage

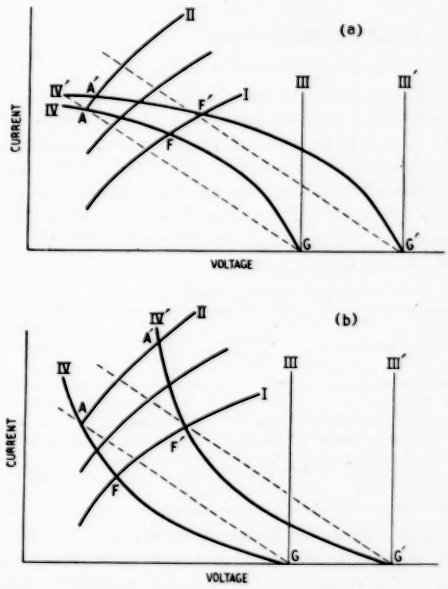

Fig. 6. *Effects of using non-linear series resistances. At (a) the resistor is a tungsten filament lamp and at (b) it is thyrite or similar material.*

by an amount which is sufficient to strike the stabilizer, even when the full load is also connected, and that standard rectifier filter components may be used for L and C. The inductance-capacitance combination also

assists in smoothing any "ripple" which may be present on the main supply.

In rectifier circuits the "no-load" voltage is often high enough to ensure striking, so that the load-switching arrangement may be used. Alternatively, the last *LC* combination in the filter circuit of the rectifier can be modified by adding resistance, obtaining its value by using for Curve III of Fig. 5 the voltage-current relation across the capacitor preceding the combination. A switch which applies the h.t. to the rectifier after its heater has reached working temperature will then give the same conditions as with d.c. mains. Such a switch may readily be placed in the lead between h.t. negative and the centre-tap of the rectifier transformer in a full-wave circuit. Where the load is of the valve type in which the heaters reach their working temperature some seconds after that of the rectifier, no switching becomes necessary.

5. Other Considerations

5.1. *Choice of stabilizer tube.*

The stabilizer tube chosen for a particular application will depend on the voltage of the source, the normal load current, and the degree of stabilization required, and the conclusions of Section 3.3 have to be taken into account. Table I indicates some of the types of tube available for currents likely to be used.

TABLE I Voltage-Stabilizer Tube Data

Designation	Running voltage	Current range	Approximate striking voltage
1. Cossor S130 ...	115-135 V	75 mA (max)	180 V
2. R.C.A. VR75-30	75 V	5-30 mA	—
3. R.C.A. VR105-30	105 V	5-30 mA	—
4. R.C.A. VR150-30	150 V	5-30 mA	—
5. R.C.A. 874 ...	90 V	5-30 mA	120 V

Since tubes of different types can be used in series, it is evident that combinations of them can be chosen to stabilize d.c. supplies whose nominal voltages cover a considerable range without having too large a difference between stabilized and nominal supply voltages.

5.2. *Safeguarding the stabilizer tube.*

It was noted in Section 3.3 (b) that a large reduction in load current might give an excessive stabilizer current. There are two methods of overcoming this difficulty. The total current drawn from the source can be maintained at its proper value without increase in stabilizer current by switching in an appropriate "dummy" load" in parallel with the main load. This can be further elaborated by having a suitable variable resistor for the dummy load, with a control calibrated so that it can be set appropriately when the equipment which forms the load is changed. The stabilizing properties of the arrangement with respect to supply voltage variations are then much the same for any load current in the range used.

A better scheme is to introduce additional resistance in series with r (or to vary r by a calibrated control) when the load current is decreased. This has the effect of accommodating the design of r to the appropriate load current, and, as shown in Section 3.3 (c), some increase in the degree of stabilization obtained is achieved by this. Since the power dissipation in the dummy load is eliminated, and that in the resistor is diminished, this arrangement decreases unwanted power consumption, whereas the first increases it.

5.3. *Some experimental results*

By way of illustration of the degree of stabilization obtainable with a comparatively small difference between stabilized voltage and nominal supply voltage, some results obtained by the procedure of Sections 3.2, 3.4, and 3.5, and checked experimentally, will be quoted. They also indicate that a value of stabilized voltage very suitable for many purposes can be obtained from d.c. mains voltages which are sometimes available, especially from lightly-loaded heavy-duty mercury-arc rectifiers, or from low-power rectifier arrangements working from a transformer secondary of about 250 V r.m.s.

The experimental procedure was as follows:—

The voltage V_2 was measured to the nearest 0.5 V with a 200 : 1 resistance potential divider and a laboratory potentiometer of the normal type. With the supply voltage at a given minimum, the resistance and the load were varied until the required load current and the minimum stabilizer current were flowing. The value of r was then noted, and the effects of changes of load current and supply voltage were investigated, and compared with those deduced

TABLE II. Experimental Results.

Stabilizer tubes :—two Cossor S.130 in series ; striking voltage 278V.

Expt. No	Nature of supply	I_2 (mA)	I_3 (mA)		V_2 (V)		r (ohms)	
			Obs.	Calc.	Obs.	Calc.	Obs.	Calc.
(i)	250 V d.c. ...	60	15	15	237.6	237.8	160	162
		30	50	47	237.2	237.5		
		0	73	68	238.6	238.8		
	260 V d.c. ...	60	79	70	239.2	239.0		
(ii)	250 V d.c. ...	30	15	15	237.8	237.8	270	271
		0	48	46	237.2	237.4		
	260 V d.c. ...	30	56	52	237.5	237.6		
		0	78	76	239.5	239.5		
(iii)	260 V d.c. ...	60	15	15	238.0	237.8	300	296
		30	49	46	237.2	237.5		
		0	74	71	239.0	239.0		
	270 V d.c. ...	60	53	50	237.6	237.6		
		30	77	74	239.4	239.2		
(iv)	260 V d.c. ...	30	15	15	237.9	237.8	500	493
		0	47	45	237.3	237.4		
	270 V d.c. ...	30	39	36	237.2	237.4		
		0	66	64	238.2	238.4		
(v)	Rectified ...	60	15	15	238.2	237.8	820	830
	225 V a.c. ...	30	46	45	237.7	237.5		
		0	75	74	239.1	239.3		
	Rectified ...	60	25	23	237.6	237.6		
	235 V a.c. ...	30	55	53	238.0	237.6		
		0	83	82	240.0	240.0		
(vi)	251.5 V d.c. ...	60	15	15	237.9	237.7	40 W, 200 V metal filament lamp used in place of r.	
		30	45	48	237.4	237.7		
		0	75	74	239.0	239.3		
	261.5 V d.c. ...	60	30	32	237.2	237.4		
		30	60	61	237.9	238.2		
		0	87	89	240.2	240.6		
	271.5 V d.c. ...	60	40	40	237.9	237.4		
		30	68	70	238.6	238.8		

theoretically, with the results shown in Table II. When r was replaced by a 40-W lamp, the supply voltage was adjusted to give the above conditions, and the voltage thus obtained was taken as the minimum for the purpose of the experiment. Since the stabilizer tube characteristic proved slightly concave towards the high-voltage side, it is not surprising that the stabilized voltages vary in both directions from those obtained with the initial conditions. Within the limits of error imposed by the conditions of the experiment and the graphical work, the results show good agreement between theory and experiment.

APPENDIX

Establishment of the results quoted in 4.2

It will be assumed that the load behaves like a resistance R, and that the value of the inductance L is independent of the current in it. Then with the previous notation, we have that (before the stabilizer strikes)

$$V = V_2 + ri_1 + L\frac{di_1}{dt} \quad .. \quad .. \quad .. \quad (1)$$

$$i_1 = C\frac{dV_2}{dt} + \frac{V_2}{R} \quad .. \quad .. \quad .. \quad .. \quad (2)$$

with the initial conditions $V_2 = \frac{dV_2}{dt} = 0$ at $t = 0$.

Solving for V_2, we obtain

$$V_2 = \frac{V}{1 + r/R}\left[1 + \frac{2\omega_2}{(4\omega_2^2 - \omega_1^2)^{\frac{1}{2}}} \cdot e^{\frac{-\omega_1 t}{2}} \sin\left\{\frac{(4\omega_2^2 - \omega_1^2)^{\frac{1}{2}}}{2}\, t + \phi\right\}\right] \quad (3)$$

where $\omega_1 = \frac{1}{CR} + \frac{r}{L}$, $\omega_2^2 = \frac{1}{LC}\left(1 + \frac{r}{R}\right)$, $\cos\phi = \frac{\omega_1}{2\omega_2}$, $(0 < \phi < \pi/2)$ and it is assumed that $4\omega_2^2 > \omega_1^2$.

Thus V_2 executes a damped oscillation about its final steady value $V\Big/\left(1 + \frac{r}{R}\right)$ with period $T = 4\pi/(4\omega_2^2 - \omega_1^2)^{\frac{1}{2}}$.

The greatest positive value of V_2 is reached

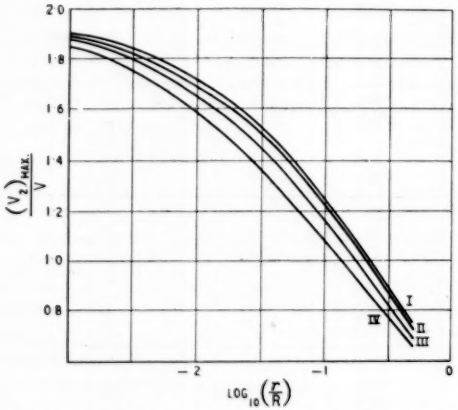

Fig. 7. *Relation between $(V_2)_{max}/V$ and r/R for different values of k ; I, $k=1$; II, $k=1/2$ or 2 ; III, $k=1/4$ or 4 ; IV, $k=1/8$ or 8.*

during the first upward rise of voltage after switching on, and is given by

$$\left(V_2\right)_{max} = \frac{V}{1 + r/R}\left[1 + \exp.\left\{-\pi\omega_1 (4\omega_2^2 - \omega_1^2)^{-\frac{1}{2}}\right\}\right] \quad (4)$$

Substituting for ω_1 and ω_2, and writing

$$\frac{L}{C} = k \,.\, r \,.\, R \quad .. \quad .. \quad .. \quad .. \quad (5)$$

equation (4) becomes

$$\left(V_2\right)_{max} = \frac{V}{1 + r/R}\left[1 + \exp.\left\{\pi\left(\frac{4k}{(1+k)^2}\left(1 + \frac{R}{r}\right) - 1\right)^{-\frac{1}{2}}\right\}\right] \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

For given values of V, r, and R, it is evident from (6) that to obtain the maximum value of V_2 it is necessary to make $k = 1$; i.e., to have

$$\frac{L}{C} = r \,.\, R \quad .. \quad .. \quad .. \quad .. \quad .. \quad (7)$$

To find whether the stabilizer tube will strike under any given set of conditions, the curves showing the relation between $(V_2)_{max}/V$ and r/R for various values of k may be used. These may be computed from (6), and are shown in Fig. 7. As long as the value of r/R is such that the ordinate to the appropriate curve exceeds the ratio (striking voltage)/V, striking will occur.

As an example, using the figures of Table II, for the 250-V d.c. supply the ratio is $\frac{278}{250} = 1.112$. From Fig. 7, the maximum values of r/R for which striking will occur are as follows :—

k	1	$\frac{1}{2}$ or 2	$\frac{1}{4}$ or 4	$\frac{1}{8}$ or 8
r/R	0.16	0.15	0.14	0.13

The choice of values for L and C is evidently a wide one in this case, since, from the first line of Table II, $r = 162\ \Omega$, $V_2 = 237.8$ V, $I_2 = 60$ mA, giving $R = 3960\ \Omega$, and $r/R = 0.041$, well within the above limits. From (7), a value of $L/C = 162 \times 3960$ and (say) $L = 10$ H, $C = 15.6\ \mu$F gives maximum upward surge ; but the wide variations permissible in the value of L/C imply that normal rectifier filter components will in general be adequate to ensure striking. Individual cases may be treated in a similar way.

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ELECTRO-ENCEPHALOGRAPH AMPLIFIER*

By Denis L. Johnston, B.Sc.(Eng.)

SUMMARY.—An amplifier with frequency response of 0.15–10,000 c/s is described, it requires a 15-μV peak-to-peak input signal for full output drive, and operates on a.c. supplies without batteries by an extension of the technique of electronically-stabilized power-supply circuits. The overall mains-frequency hum signal is equivalent to less than 0.5-μV peak-to-peak input signal. Four channels of amplification and an ink-writing recorder are assembled in a compact equipment for electro-encephalography. Some special properties and unusual circuit arrangements of differential amplifiers are discussed. Critical balancing adjustments have been eliminated by the use of degenerative circuits.

1. GENERAL DESCRIPTION
1.1. Discussion of Requirements.
1.2. Mechanical Arrangement.
1.3. Electrical Arrangement.

2. AMPLIFIER CIRCUIT
2.1. *Input Stages*
2.1.1. Coupled-cathode differential amplifier.
2.1.2. Differential input circuit.
2.1.3. Differential amplifier with floating h.t. supply.
2.1.4. Valve and circuit noise at very low frequencies.
2.1.5. Electromagnetic and electrostatic induction.
2.1.6. Other causes of spurious fluctuations.
2.2. *Intermediate Stages*
2.2.1. Basic circuit.
2.2.2. Low-frequency response.
2.2.3. High-frequency response.
2.2.4. Reduction of blocking time.
2.2.5. Feedback in differential amplifiers.
2.3. *Output Stage*
2.3.1. Direct-coupled drive circuit.
2.3.2. Recorder drive arrangements.

3. ELECTRONICALLY-STABILIZED SUPPLIES
3.1. Discussion of Performance.
3.2. Requirements of H.T. Stabilizing Circuit.
3.3. Design of H.T. Stabilizing Circuit.
3.4. Power Unit Chassis.
3.5. Stabilized Heater Supply.

1. General Description.

1.1. *Discussion of Requirements*

A VERY complete survey of the medical and instrumental practice in electro-encephalography (e.e.g.) has been given by Parr and Grey Walter,[1] and designs of special amplifiers described by a number of authors[2–9]. It has always been found necessary to use batteries for the high-tension supply to the earlier stages, and frequently for the heater supplies also.

The frequency range required for the graphic recording of brain potentials is at least 0.3–70 c/s. The lower limit is commonly set by problems of amplifier stability, and the upper limit by the available direct-writing recorders. As mains-supply frequencies fall within this range it is not easy to attain a sufficiently low hum level. Electro-encephalographic potentials are not reproduced satisfactorily if extraneous noise and hum exceeds a level equivalent to about 2-μV peak-to-peak input signal.

* MS. accepted by the Editor, October 1946.

In certain equipments[5] mains-frequency components have been eliminated by a low-pass filter, but this solution is hardly adequate with 60-c/s supplies, and with 50-c/s supplies a cut-off placed at about 40 c/s obscures important components of the e.e.g. potential waveforms. An alternative approach has been the use of carrier frequency and chopper-type amplifiers[10–13], but this method introduces a higher level of noise.

It is usual to use accumulators or other batteries to supply or stabilize the voltage to the heaters of valves in early stages in order to reduce drift in the amplifier. In the equipment to be described such batteries are eliminated. This self-contained equipment does not require the special screening and spacing previously necessary in clinical use[14], and no attention has to be paid to the maintenance of accumulators and dry batteries.

1.2. *Mechanical Arrangement*

It was found possible to arrange the amplifier and power-unit chassis in close

proximity without appreciable induction of mains-frequency voltages. The chassis are to the dimensions of the 19-inch standard rack and are mounted in a steel cabinet 40 inches high fitted with castors. A rear door carries the power units, and when this is lowered the valves of all chassis are easily reached (Fig. 1). The complete equipment incorporates four channels of amplification in two two-channel units.

Access to wiring and components of the amplifiers is obtained by opening the hinged front panels of the chassis (Fig. 2). Figure 3 shows the complete electro-encephalograph equipment in use : a section of the cabinet side lifts to form a working desk that is level with the recorder set in the top of the cabinet of Fig. 4, which shows the first experimental model.

It was necessary to take some precautions against microphony from vibration of the motor driving the paper tape for the recorder. The motor and paper drive unit are mounted on rubber anti-vibration units, and the critical stages of the amplifier are on rubber-mounted subchassis. The "click" springs on rotary switches are adjusted to be only strong enough for correct location.

1.3. *Electrical Arrangement*

Differential-input amplifiers are employed in electro-encephalography because the potential to be observed is that between two points on the scalp rather than between one

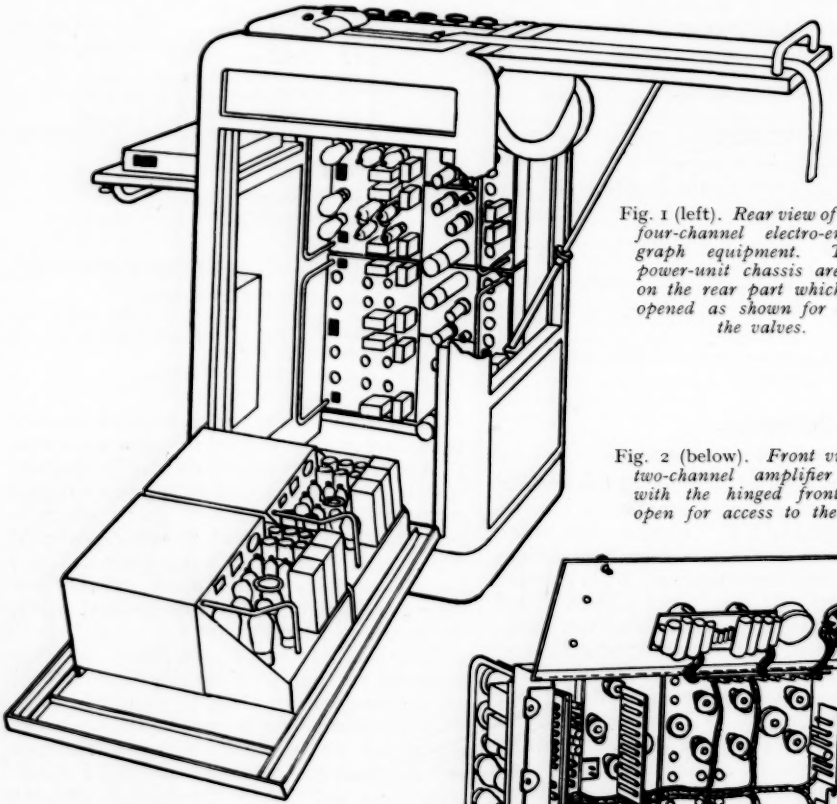
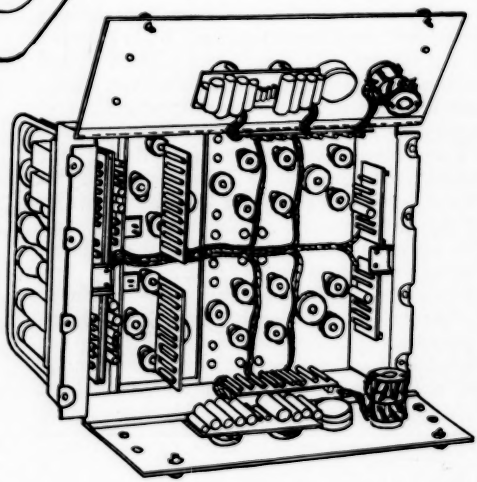


Fig. 1 (left). *Rear view of complete four-channel electro-encephalograph equipment. The two power-unit chassis are carried on the rear part which can be opened as shown for access to the valves.*

Fig. 2 (below). *Front view of a two-channel amplifier chassis with the hinged front panels open for access to the wiring.*



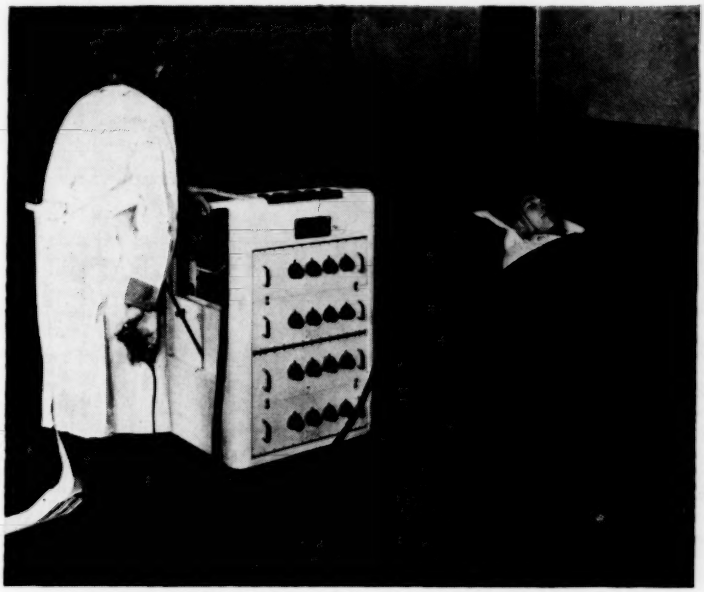


Fig. 3. *Experimental model of electro-encephalograph in use.*

point and an arbitrary "earth." These special requirements are explained by Parr and Grey Walter[1] and others[15–17]. The screening of the patient from external electrical interference is made very much easier if the differential amplifiers can discriminate against in-phase inputs; i.e., potentials appearing equally between each point on the scalp and earth.

In the present design it has been possible without critical adjustment to obtain a discrimination of as high as 3,000 : 1 between the amplification of differential and of in-phase input signals, by using a cascade of coupled-cathode push-pull stages of amplification. For maximum effect the cathode-coupling load at the first stage is provided by the dynamic anode impedance of a pentode.

As the early stages have to be push-pull and a push-pull output is desired, it is convenient to make the amplifier push-pull throughout. This confers certain advantages and it facilitates the provision of gain controls which are placed differentially across anode or cathode loads. The high-frequency response control is a switched differential capacitor and the low-frequency control a pair of switched series coupling capacitors.

It has been pointed out by Dawson[18] that the reproduction of electro-cardiograph and electro-encephalograph waveforms can be seriously modified by the phase shift in

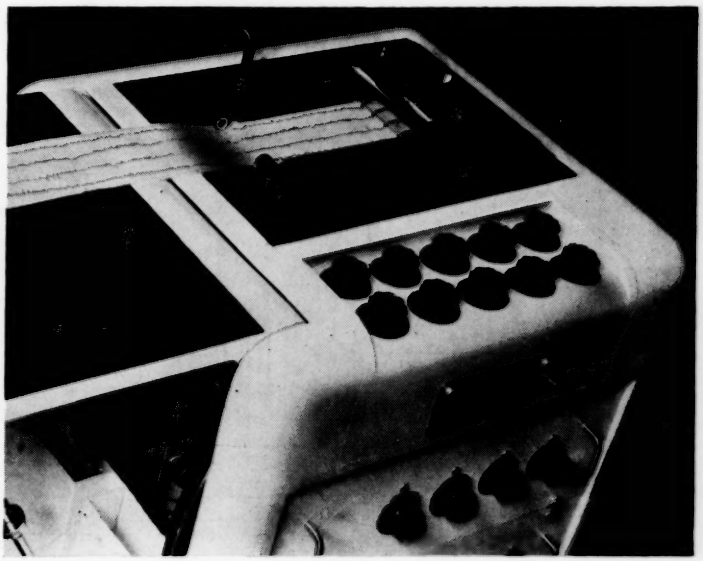


Fig. 4. *Recorder and head-electrode selector switches in desk-top of complete equipment.*

several capacitor-coupled stages. A direct-current amplifier is not satisfactory because relatively large and slowly-changing muscle potentials would mask the desired record. The ideal amplifier has one capacitor-coupled stage. In the present design, of the four interstage couplings, two are direct. The small number of capacitor couplings is of assistance in obtaining the relatively long time constant of one second, without the use of unusually large coupling capacitors and grid leaks.

A schematic circuit of the amplifier is given in Fig. 5.

A high degree of stabilization is required of the power-supply circuits and for the sake of economy in a multichannel equipment one power unit is arranged to supply a pair of amplifier channels.

The high-tension supply must duplicate the characteristics of a dry-battery that is insulated and "floating" to earth. A block schematic of the stabilizer arrangements is shown in Fig. 6. For the h.t. supply two stages of a conventional electronic-stabilizer circuit are used in series, and a third stage of stabilization causes the h.t. supply to "float" at a definite potential to earth. These last two stages are not on the power unit chassis but are located physically adjacent to the input stages of the amplifier chassis.

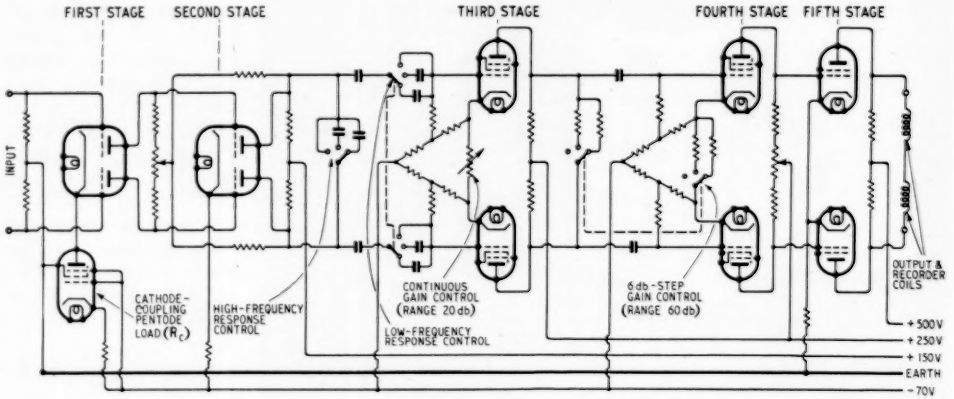


Fig. 5. *Simplified basic circuit of complete differential amplifier.*

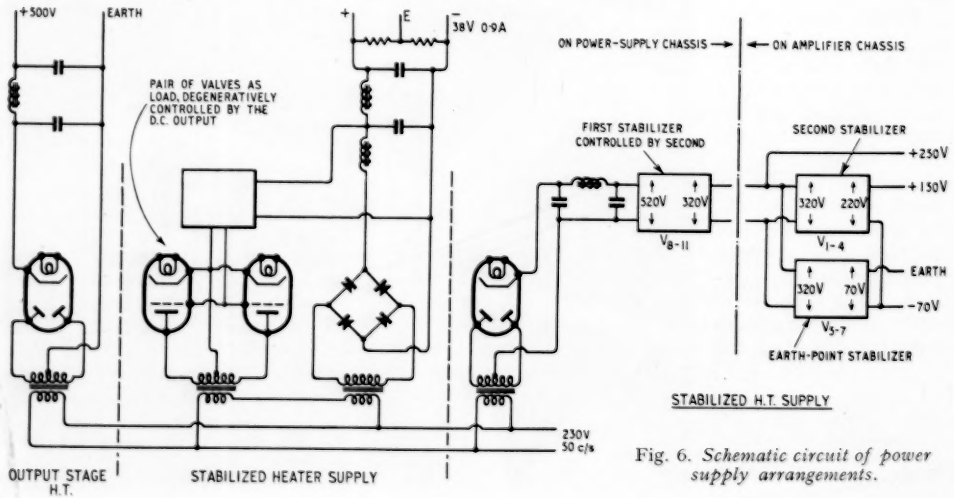


Fig. 6. *Schematic circuit of power supply arrangements.*

Controlled voltage for the heater supply to the critical stages is derived from an electronic a.c. stabilizer followed by a metal rectifier and smoothing circuit. Current for the output stage of the amplifier is supplied by a rectifier and smoothing circuit without stabilization.

2. Amplifier Circuit

2.1. *Input Stages*

2.1.1. *Coupled-cathode differential amplifiers*

The term "coupled-cathode differential amplifier" is used here rather than the customary term "cathode-coupled" as the latter suggests inter-stage coupling, and we wish to imply a coupling between the cathodes of a pair of valves forming a single push-pull stage as shown in Fig. 7(*a*). The amplification of this stage to push-pull signals is: (taken differentially)

$$M' = \frac{\mu \,.\, R_L}{r_a + R_L}$$

and the amplification to in-phase signals: (taken at either anode load)

$$\overline{M} = \frac{\mu \,.R_L}{r_a + R_L + 2R_c(\mu + 1)}$$

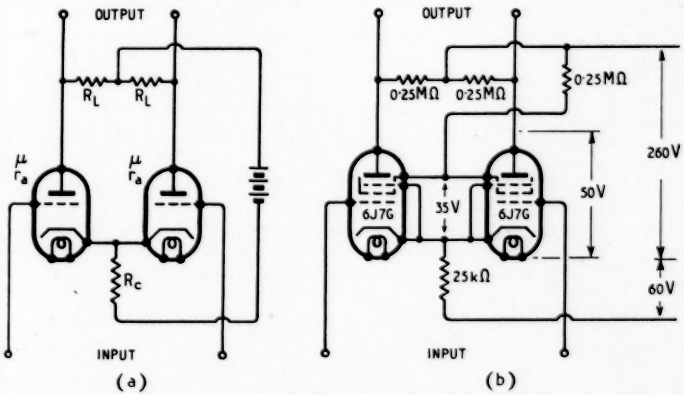


Fig. 7. *Coupled-cathode amplifier stage (a) and practical circuit with values for 6J7G pentode valves (b).*

For an amplifier using 6J7G pentodes with the circuit of Fig. 7(*b*) the values of amplification M' and $\overline{M}$ are respectively 140 and 5. If the valves are connected as triodes the amplifications are respectively 16 and 4.

We see that the discrimination factor F is given by:—

$$F = \frac{M'}{\overline{M}} = \frac{r_a + R_L + 2R_c(\mu + 1)}{\mu \,.\, R_L}$$

and in the above circuits the value is 28 for the pentodes and 4 for the triodes.

When the amplifier is employed as a phase

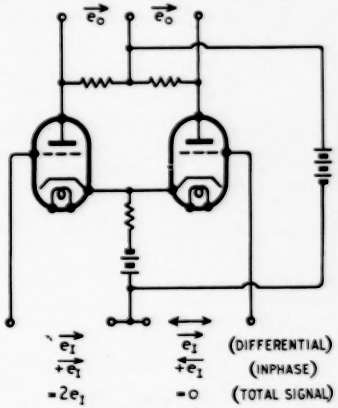


Fig. 8. *Self-balanced phase invertor.*

invertor (Fig. 8) the signal between one grid and earth is zero and the input is applied between the other grid and earth. This is the equivalent of superposing a push-pull signal of $2e$ across the pair of grids, and a signal of e in-phase between each grid and earth.

D is:

$$D = \frac{\text{In-phase signal output}}{\text{Push-pull signal output}} = \frac{\mu \,.\, R_L}{r_a + R_L + 2R_c(\mu + 1)}$$

Thus the inversion defect in a degenerative phase-invertor is the reciprocal of the discrimination factor to in-phase signals.

Balanced phase-invertors and composite balanced and degenerate circuits are in common use: various means are adopted for balancing out the phase inversion defect, but they are appropriate only to constant-input circuit conditions and for this reason are unsuitable in the detection of brain potentials.

The differential amplifier circuit was first applied by a number of independent workers

between 1936 and 1938[3,19–24]. Offner pointed out the high stability possible with the circuit, and Goldberg has recently contributed the use of a pentode valve as the coupling inpedance[4]. The coupled-cathode circuit has been analysed and discussed by a number of the above and by other workers[26,27]. Special applications such as oscillator circuits have been described recently[28–32] and the earlier work that was carried out with balanced-differential circuits has been summarized by Penick[61] and others[34–37].

Miller used twin-triode valves in a coupled-cathode amplifier to obtain a high degree of stability in what was functionally a single-sided amplifier[39]. Here one half of the twin triode was regarded as the amplifier and the other half treated as a degenerative stabilizer, the circuit having the form shown in Fig. 9.

This form of circuit can similarly improve the stability of a differential amplifier, and

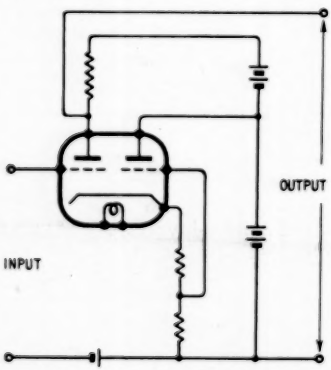


Fig. 9. *Basic circuit of Miller's amplifier.*

for the performance required of the present design, its use was found essential, both for the low-level stages of amplification and for the electronic-stabilizer circuits. The circuit discriminates against variations in high-tension supply and heater voltage just as it rejects in-phase grid voltages.

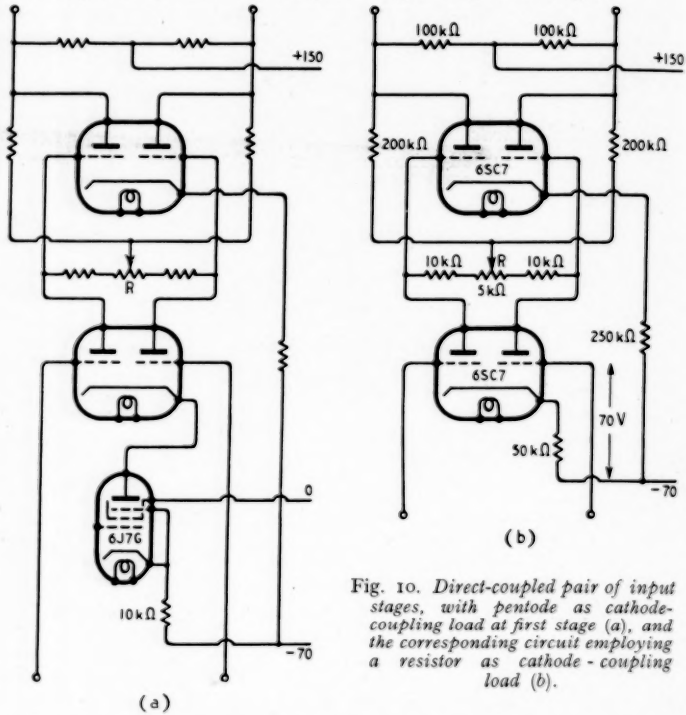


Fig. 10. *Direct-coupled pair of input stages, with pentode as cathode-coupling load at first stage (a), and the corresponding circuit employing a resistor as cathode-coupling load (b).*

2.1.2. *Differential Input Circuit.*

The discrimination factor of the amplifier shown in Fig. 7(*b*) is 28. If it is backed up by a following stage that also possesses a high rejection factor, the overall rejection is higher than that of a single stage. In the present equipment the first two stages are direct coupled and triodes are used [Fig. 10(*a*)]. A high degree of negative feedback of in-phase voltage is applied, principally to minimize the effect of h.t. variations. A limitation in cascaded direct-coupled amplifiers is the available high-tension voltage. In Fig. 10(*b*) employing 6SC7 triodes a discrimination factor of 69 is obtained when the cathode load is 50,000 ohms and the p.d. across the load is 70 volts. It is uneconomical to increase the discrimination factor substantially by increasing the cathode-load resistor, but it is possible to use a pentode in this position as in Fig. 10(*a*). With a 6J7G pentode a dynamic impedance

of 5 MΩ is obtained with this same potential drop of 70 V. The value of the dynamic impedance r'_a for a pentode with self-bias is given by $r'_a = r_a + (\mu + 1)R'_c$ approximately: it is reduced by about 10 per cent in the 6J7G by the mutual conductance between the screen grid and anode.

The discrimination factor is increased in proportion to the value of the cathode load. If adjustments are made to compensate for the inequalities of characteristics in a pair of valves with differential input and output, it is possible to realise a factor of 100,000 : 1; using components of commercial tolerance 3,000 : 1 is obtained in the circuit of Fig. 10(*a*) without further adjustment.

The performance can be improved further by taking feedback to the load-valve grid[60] as shown in Fig. 11 for this case of triodes, but in the circuit adopted this was not done because readjustment becomes necessary when valves are changed.

The limit of performance of the differential

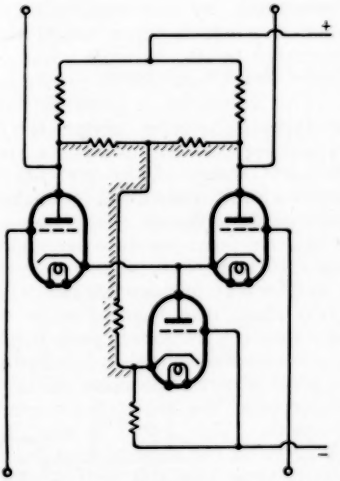


Fig. 11. *Differential input amplifier with feedback (cross-hatched) to grid of valve serving as dynamic cathode load.*

amplifier is reached when the level of the in-phase signal is high enough to swing the valve characteristic outside the linear range: intermodulation then occurs between the in-phase and the push-pull components. This limiting level can be raised by applying negative feedback individually to each valve (Fig. 12).

2.1.3. *Differential Amplifier with Floating H.T. Supply.*

The differential amplifier is intended to work with its input grids at about earth potential, and when an h.t. battery is used, an earth tap must be provided at a suitable

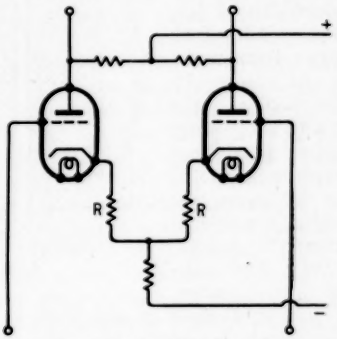


Fig. 12. *Differential input circuit for high signal levels.*

potential between the negative and positive poles. An earth point can be taken on a potentiometer connected across the battery, but the impedance will be relatively high and careful screening against electrical interference is then necessary. This difficulty will also arise with a mains-derived power supply, and the method of overcoming it is described in Section 3.1.1.

2.1.4. *Valve and Circuit Noise at Very Low Frequencies.*

The chief sources of spurious output voltages will be:

(*a*) Thermal agitation noise of the input circuit.

(*b*) Shot-effect noise of the first stage valves.

(*c*) Noise due to space charge variations of a random nature, and variations induced by changes in heater current.

(*d*) Modulation at the several electrodes of valves in the early stages due to fluctuations in high-tension voltage.

(*e*) Other causes, such as mechanical vibration and electromagnetic and electrostatic induction.

In a differential amplifier a source peculiar to one side of the amplifier will be amplified as a differential signal: as Pratt[40] has

pointed out, the application of negative feedback has no effect on the equivalent input noise of an amplifier.

The relative importance of the several sources of noise will depend upon the frequency range and the bandwidth of the amplifier. The first two are the main sources of noise in radio-frequency amplifiers and they have received a good deal of attention: useful reviews are given by Pearson[41], Thompson[42] and others[5,43].

At very low sub-audible frequencies, circuit and shot noise is not high because the bandwidth is small, and the greater part of the fluctuations appear to originate at the cathode space charge, and fall below about 10 c/s. These effects are least in large triodes of low amplification factor. The condition of the space charge depends to some extent upon the heater current, and consequently the heater supply must be stablized in sensitive low-frequency amplifiers: the performance of indirectly-heated valves is better than directly-heated types in this respect because of the thermal inertia of the indirectly-heated cathode.

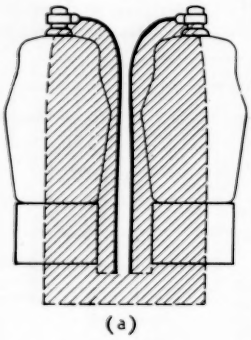


Fig. 14. *Effective loop areas for induction of stray electromagnetic fields in typical valves; (a) two type-6J7G, (b) two type-6SJ7, (c) twin-triode, type-6SC7.*

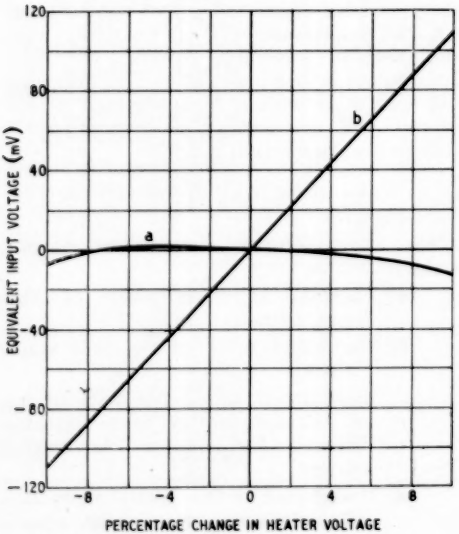


Fig. 13. *Fluctuation due to changes in heater voltage; (a) in Miller's circuit, and (b) in conventional single-sided amplifier.*

Thompson gives an expression for thermal agitation hoise in a circuit of resistance R $e = \sqrt{1.6 \times 10^{-20} \times R \times \Delta f}$ where Δf is the bandwidth. The shot effect noise can be represented by an equivalent noise resistance R_{eq}, which for a triode is given approximately by $R_{eq} = 2.5/g_m$, and which for a type 6SC7 twin triode valve is 1900 ohms. The value for a pentode valve of similar type is usually about ten times higher, and for this reason triodes are used for the input stages of the present design. The input circuit resistance in e.e.g. work is usually of the order of 10,000 ohms. The bandwidth of an electro-encephalograph working into a mechanical recorder may be 60 c/s, and for an equivalent circuit resistance of 10,000 ohms, the thermal-noise voltage will be about 0.3 microvolt peak-to-peak.

The noise level observed at low frequencies with a good selected specimen of the 6SC7 type valve is of the order of 1.5 microvolts peak-to-peak. This valve is used for the reasons given in the next section, because it is the only type available with a physically common cathode. The single triode MH40[45] is a type of valve known to have a noise value about half the above figure but did not meet all the requirements of a self-contained equipment. There is little published information on low-frequency noise, but Roess[46] has given some useful data taken at very narrow bandwidths in connection with infra-red spectroscopy.

Miller[39] has shown the reduction in drift with change in heater voltage that can be realised by using a balanced-cathode circuit,

in the curve (*a*) of Fig. 13 reproduced from his paper: (*b*) is the performance of a conventional amplifier. The improvement is greatest when the valve used is a twin-triode with a cathode physically common to both triode sections (such as the R.C.A. 12SC7 or 6SC7). As might be expected, the changes in cathode conditions due to variation in heater current are then more nearly equal in the two sections than they would be in individual valves, and being equal the in-phase discrimination of the amplifier comes into effect when such valves are applied in differential circuits.

Some work on these lines was done a few years ago by Brentano[35,36] with a valve designed so that the two triode sections received emission from substantially the same area of the filament. An American miniature indirectly-heated double triode, the 6J6, has a flat cathode with the triode sections on either side,[25]. Unfortunately specimens of these two types of valves have not been available for trial, for they are the only other known types particularly suited to the present application.

2.1.5. *Electromagnetic and Electrostatic Induction*

At low signal levels differential pickup of electromagnetic fields is experienced within the circuit wiring. It can be reduced by keeping at a minimum the area of pickup loop formed by the differential circuit, and by a degree of magnetic screening. Magnetic induction can be minimized by choosing the most suitable valves. The worst arrangement is a pair of top-grid pentodes, where the pickup loop area is large. Fig. 14(*a*) represents a pair of 6J7G valves; the loop area is about 60 square cm for valves and circuit wiring. In an alternating field of one gauss (peak) the voltage picked up at 50 c/s would be 30 microvolts (peak).

An under-grid valve such as the 6SJ7 is better Fig. 14(*b*), the loop area being about 15 square cm. The best valve available is the 6SC7: the two triode sections are above one another on the same vertical axis and the total area between the pair of grid leads and wiring is about 3 square cm. This type of valve is enclosed in a ferrous envelope and there is considerable magnetic screening, particularly when the valve holder is fitted in an enclosed steel chassis.

Electrostatic induction is not troublesome in the present instance since all wiring to the first two stages is screened by the chassis, and input cables are screened and fitted to screened plugs and sockets. Pairs of signal leads are run close together in the amplifier in order that the pickup shall be as nearly as possible equal on each lead and so be rejected by the differential amplifier.

2.1.6. *Other Causes of Spurious Fluctuation*

The insulation resistance between heater and cathode in small indirectly-heated valves may not be very much greater than the order of one megohm. In a single-cathode twin-triode circuit the effect of heater-cathode leakage is reduced considerably because the voltage due to leakage is common to each side of the amplifier. In a sensitive input

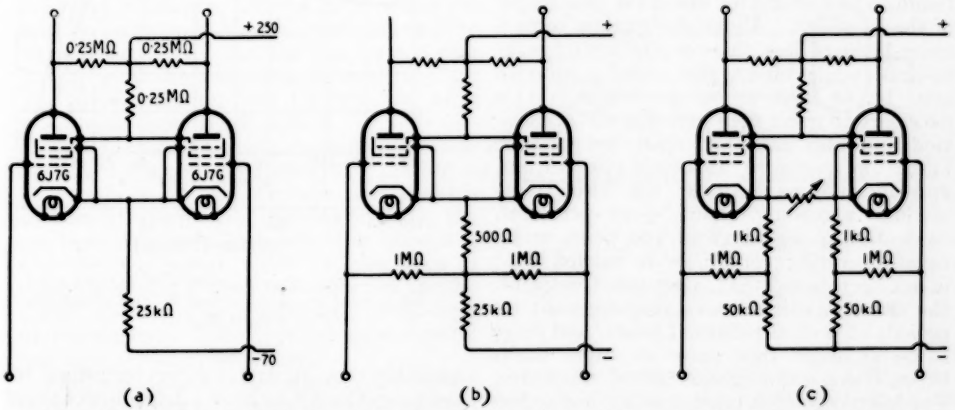


Fig. 15. (*a*) *Basic circuit for 3rd and 4th stages;* (*b*) *circuit with self-bias connections;* and (*c*) *alternative self-bias circuit with cathode gain control.*

circuit leakage current is reduced by arranging that the cathode-heater potential difference is small: this is analogous to the use of a guard electrode in measurement technique.

Grid leaks are connected at the input of the amplifier. The value is one megohm, and as this is rather high for a wire-wound

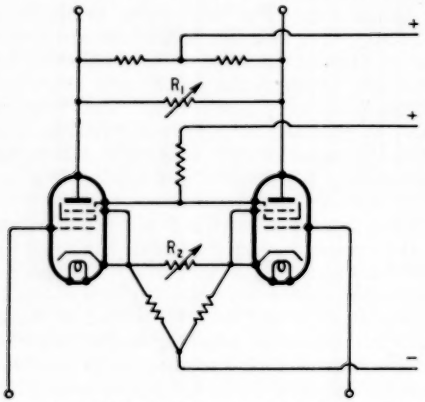


Fig. 16. *Differential gain control in anode load circuit R_1 and in cathode load circuit R_2.*

component, cracked-carbon film resistors are used. It is essential that the grid current be of very low order of magnitude or there will be generation of noise in the resistors. The other resistors in the first two stages are such values that wire-wound components can be used in every case.

Only a proportion of commercial valves are found to be suitable for use in the first stages of the amplifier. About one quarter show a noise level of less than 2 microvolts peak-to-peak while about the same proportion may be as high as 20 microvolts. It is necessary to reject such valves in this application and also any that show low cathode insulation resistance, abnormal grid current, and inequalities between the two triode sections exceeding about $\pm$ 10 per cent. An initial ageing of about 100 hours under rated operating conditions is carried out; it has been found that, after this treatment, the characteristics remain very constant for periods of several thousand hours, and there is no evidence that noise or other faults become worse during the life of the valve. The initial ageing is important for any valves that are in directly-coupled circuits: during the first few hours the base voltage of the grid characteristic appears to shift, making it impossible to set up the circuit properly.

Microphony gives very little trouble in valves of modern construction, and the all-metal type 6SC7 is especially good in this respect. The complete equipment must be struck with a fairly heavy blow to produce a substantial deflexion on the recorder when the amplifiers are working at maximum sensitivity.

The natural frequencies of valve structures fall in the range 100–10,000 c/s. Resiliently mounted sub-chassis are employed for the early stages of the equipment. They are fitted with small rubber-metal bonded units of nominal loading one pound. The natural frequency of the sub-panel assembly is thereby reduced to the range 1–10 c/s to which frequencies the valve structures present relatively high stiffness.

A resiliently mounted assembly will have a natural frequency given by the expression $f = \sqrt{10/h}$ c/s, where h is the deflection of the mountings in inches due to the weight of the

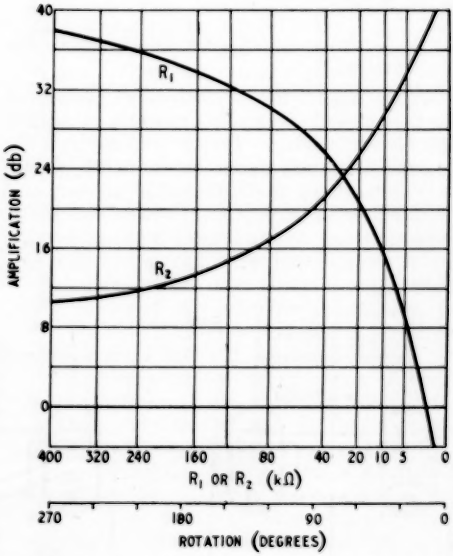


Fig. 17. *Law of gain controls of tapered track when used in anode circuit R_1 and in cathode circuit R_2.*

assembly. A typical rubber mounting for one pound load may have a deflection of about 1/8 inch, corresponding to a natural frequency when loaded of 9 c/s.

2.2 *Intermediate Stages*

2.2.1 *Basic Circuit*

The third and fourth stages of the amplifier contribute the major part of the total voltage gain of over 140 db, the gain per stage being rather more than 40 db. Type 6J7G pentodes are employed in the circuit of Fig. 15(*a*). The anode current is approximately 0.7 mA

easily demonstrated by using the star-delta transformation theorem as shown in Fig. 18 (*a*) and (*b*). A simpler means of estimating the amplification to differential signals is to assume that the cathode or anode load effective to differential signals is

$\frac{1}{2}$. R_2 in parallel with R_1 at the anode and
$\frac{1}{2}$. R_4 in parallel with R_3 at the cathode

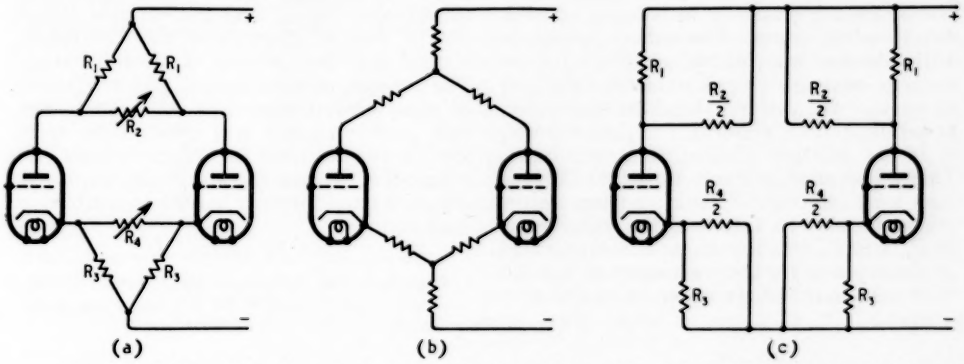


Fig. 18. *Differential gain controls at anode and cathode represented as "delta" networks (a), and as "stars" (b). The equivalent circuit of differential loads to push-pull signals (c).*

which is lower than usual practice but provides about the optimum ratio of gain to overall high-tension voltage. Self bias can be applied in this circuit as in Fig. 15(*b*) or (*c*) with no appreciable loss of amplification.

Pairs of ganged controls are undesirable for practical reasons. In a differential amplifier gain controls acting on the anode and on the cathode circuits can be placed differentially as in Fig. 16. The anode control R_1 increases amplification with n-crease of resistance. At the cathode, amplification increases as the variable resistor R_2 is reduced in value, the law for the case of the circuit Fig. 16 being given in Fig. 17. The shape is improved by the use of a 5 : 1 linear taper track control.

In the present design a continuously variable gain control is required : by placing it as a cathode-coupling control, it is possible to use a wire-wound component. In the position differentially across the anodes a considerably higher value of resistance is required, and it is then necessary to use the rather less desirable composition track type of component. This type of gain control was first reported by Jofeh[38].

The cathode and anode attenuation control circuits are delta networks, and their relation to the more familiar circuit arrangement is as in Fig. 18(*c*). This follows from the fact that with a push-pull signal the mid-points of R_2 and R_4 are at zero signal potential

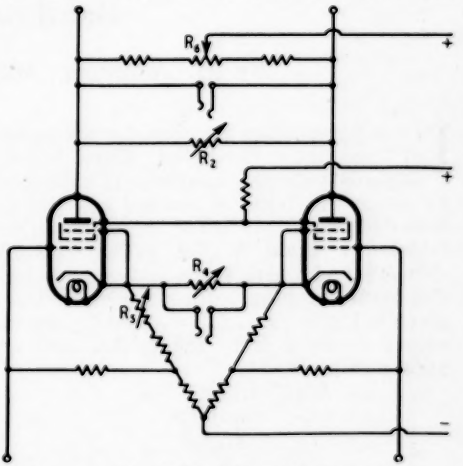


Fig. 19. *Static balancing controls R_5 at cathode and R_6 at anode, with cathode and anode test jacks.*

relative to the negative pole, and so are effectively connected together and to the negative pole. The anode and cathode gain

controls are independent of one another. Amplification of in-phase signals is constant for any position of the cathode control, and will vary with the value of the anode differential resistor.

An undesirable surge voltage occurs as the differential control is operated, in either the anode or cathode positions, when the potentials of the pair of valves at these points are not statically in balance at their direct current values. The cathode potentials are made exactly equal by including a pre-set variable resistor of small value R_5 (Fig. 19) as part of the self-bias load for one valve. It is adjusted for zero cathode p.d. when R_4 is in its position of maximum resistance. This adjustment requires attention only at very long intervals of time, or when valves are changed: to facilitate the adjustment of R_5 and R_6 small tests jacks are provided at these points for the connection of a meter with sensitivity of the order of 20,000 Ω/v.

The present amplifier design requires a continuous attenuation control of 20-db range, and a variable control with 6-db steps and a total range of 60 db. This exceeds the amplification of any one stage and to avoid using a screened attenuator the 60-db control is distributed over two stages of the amplifier. The 6-db steps are taken alternatively from the anode circuit of the third stage, Fig. 5, and the cathode circuit of the fourth stage, and the continuous control is at the cathode of the third stage. The resistor manufacturers' preferred series of component values is a logarithmic one, and some economy was effected by using the ± 10 per cent series to obtain the component values for the 6-db steps (the effective anode load is halved or doubled at each step).

(To be concluded).

References will appear at the end of the last instalment.

FILTER DESIGN TABLES BASED ON PREFERRED NUMBERS*

Band-pass Filters

By H. Jefferson, M.A., A.Inst.P., A.M.I.E.E.

IN this paper tables are given for the design of constant-k band-pass filters having preferred values of capacitance. The method by which this result is reached has already been described.[1] As before, the values in the tables are those of the ladder network elements and where single sections are used the actual element values used are those given in Fig. 1. L_{1k}, L_{2k}, C_{1k} and C_{2k} are the values obtained from the tables, and are approximations to

$$L_{1k} = R/\pi(f_2-f_1)$$
$$L_{2k} = (f_2-f_1)R/4\pi f_1 f_2$$
$$C_{1k} = (f_2-f_1)/4\pi f_1 f_2 R$$
$$C_{2k} = 1/\pi(f_2-f_1)R$$

Where f_1, f_2 are the cut-off frequencies and R is the impedance. These expressions take a more convenient form if we write $(f_2-f_1) = f_b$, the band-width, and $(f_1 f_2)^{\frac{1}{2}} = f_m$, the geometric mid-band frequency. Then

$$L_{1k} = R/\pi f_b \qquad C_{1k} = f_b/4\pi f_m^2 R$$
$$L_{2k} = f_b R/4\pi f_m^2 \qquad C_{2k} = 1/\pi f_b R.$$

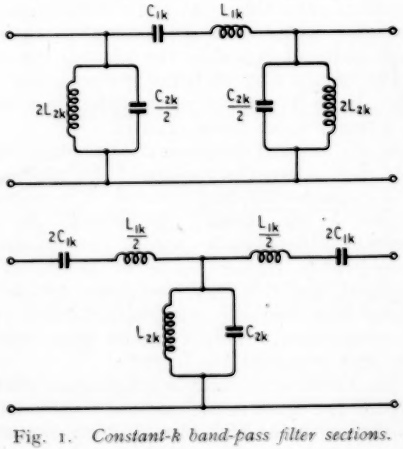


Fig. 1. *Constant-k band-pass filter sections.*

*MS. accepted by the Editor, August 1946.

It will be seen that L_{1k} and C_{2k} are the same as for a low-pass filter of the same impedance R and cut-off frequency f_b. Thus the tables already published[2] may be used to determine these two element values.

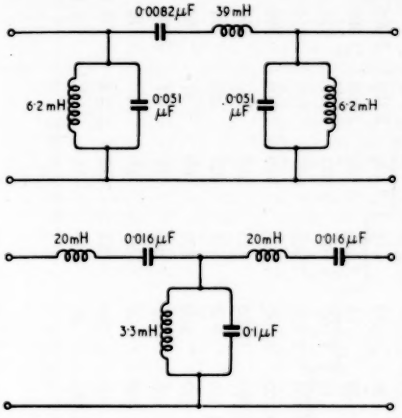


Fig. 2. *The component values of the example discussed in the text are marked on the π- and T-sections.*

We also have

$$4\pi^2 f_m^2 L_{1k} C_{1k} = 1$$

and $$4\pi^2 f_m^2 L_{2k} C_{2k} = 1.$$

Having determined L_{1k} and C_{2k} from the low-pass filter tables[2] the tables which follow can be used to determine C_{1k} and L_{2k}. Table I gives the geometric mid-band frequency f_m; this table may only be used when f_1 and f_2 are in the same decade. If f_2 is in the next decade to f_1, the value of f_m derived from Table I is taken in the top row of the supplementary table, and the true value of f_m read off below.

The values of L_{1k} and C_{2k} are obtained from the low-pass filter tables. Table II is a coarse table which enables an approximate value to be obtained for C_{1k} and L_{2k}. Table III enables the significant figures to be determined. Any practical filter will involve the use of either $2X$ or $X/2$, where

TABLE I

GEOMETRICAL MID-BAND FREQUENCY $fm = (f_1 f_2)^{1/2}$

CUT-OFF FREQUENCY f_2

CUT-OFF FREQUENCY f_1	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	100
10	10		11		12		13		15		16		18		20		22		24		27		30		33
11		11		12		13		15		16		18		20		22		24		27		30		33	
12	11		12		13		15		16		18		20		22		24		27		30		33		36
13		12		13		15		16		18		20		22		24		27		30		33		36	
15	12		13		15		16		18		20		22		24		27		30		33		36		39
16		13		15		16		18		20		22		24		27		30		33		36		39	
18	13		15		16		18		20		22		24		27		30		33		36		39		43
20		15		16		18		20		22		24		27		30		33		36		39		43	
22	15		16		18		20		22		24		27		30		33		36		39		43		47
24		16		18		20		22		24		27		30		33		36		39		43		47	
27	16		18		20		22		24		27		30		33		36		39		43		47		51
30		18		20		22		24		27		30		33		36		39		43		47		51	
33	18		20		22		24		27		30		33		36		39		43		47		51		56
36		20		22		24		27		30		33		36		39		43		47		51		56	
39	20		22		24		27		30		33		36		39		43		47		51		56		62
43		22		24		27		30		33		36		39		43		47		51		56		62	
47	22		24		27		30		33		36		39		43		47		51		56		62		68
51		24		27		30		33		36		39		43		47		51		56		62		68	
56	24		27		30		33		36		39		43		47		51		56		62		68		75
62		27		30		33		36		39		43		47		51		56		62		68		75	
68	27		30		33		36		39		43		47		51		56		62		68		75		82
75		30		33		36		39		43		47		51		56		62		68		75		82	
82	30		33		36		39		43		47		51		56		62		68		75		82		91
91		33		36		39		43		47		51		56		62		68		75		82		91	
100	33		36		39		43		47		51		56		62		68		75		82		91		100

SUPPLEMENTARY TABLE FOR USE WHEN f_1 and f_2 ARE NOT IN THE SAME DECADE

10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	100
33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33

TABLE II

INDUCTANCE OR CAPACITANCE OF BAND-PASS FILTER

INDUCTANCE		CAPACITANCE														
		μμF							μF							
		1	3.3	10	33	100	330	1 000	0.0033	0.01	0.033	0.1	0.33	1	3.3	10
	1	5 100	3 000	1 600	910	510	300	160	91	51	30	16	9.1	5.1	3.0	1.6 Mc/s
	3.3	3 000	1 600	910	510	300	160	91	51	30	16	9.1	5.1	3.0	1.6	910
	10	1 600	910	510	300	160	91	51	30	16	9.1	5.1	3.0	1.6	910	510
μH	33	910	510	300	160	91	51	30	16	9.1	5.1	3.0	1.6	910	510	300
	100	510	300	160	91	51	30	16	9.1	5.1	3.0	1.6	910	510	300	160 kc/s
	330	300	160	91	51	30	16	9.1	5.1	3.0	1.6	910	510	300	160	91
	1 000	160	91	51	30	16	9.1	5.1	3.0	1.6	910	510	300	160	91	51
	0.0033	91	51	30	16	9.1	5.1	3.0	1.6	910	510	300	160	91	51	30
	0.01	51	30	16	9.1	5.1	3.0	1.6	910	510	300	160	91	51	30	16
	0.033	30	16	9.1	5.1	3.0	1.6	910	510	300	160	91	51	30	16	9 100
	0.1	16	9.1	5.1	3.0	1.6	910	510	300	160	91	51	30	16	9 100	5 100
	0.33	9.1	5.1	3.0	1.6	910	510	300	160	91	51	30	16	9 100	5 100	3 000
	1	5.1	3.0	1.6	910	510	300	160	91	51	30	16	9 100	5 100	3 000	1 600
mH	3.3	3.0	1.6	910	510	300	160	91	51	30	16	9 100	5 100	3 000	1 600	910
	10	1.6	910	510	300	160	91	51	30	16	9 100	5 100	3 000	1 600	910	510 c/s
	33	910	510	300	160	91	51	30	16	9 100	5 100	3 000	1 600	910	510	300
	100	510	300	160	91	51	30	16	9 100	5 100	3 000	1 600	910	510	300	160
	330	300	160	91	51	30	16	9 100	5 100	3 000	1 600	910	510	300	160	91
	1	160	91	51	30	16	9 100	5 100	3 000	1 600	910	510	300	160	91	51
	3.3	91	51	30	16	9 100	5 100	3 000	1 600	910	510	300	160	91	51	30
H	10	51	30	16	9 100	5 100	3 000	1 600	900	510	300	160	91	51	30	16
	33	30	16	9 100	5 100	3 000	1 600	910	510	300	160	91	51	30	16	9.1
	100	16	9 100	5 100	3 000	1 600	910	510	300	160	91	51	30	16	9.1	5.1

TABLE III

SIGNIFICANT FIGURES FOR INDUCTANCE OR CAPACITANCE OF BAND-PASS FILTER

MID-BAND FREQUENCY f_m

CAPACITANCE OR INDUCTANCE	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	100
10	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27
11	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24
12	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22
13	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20
15	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18
16	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16
18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15
20	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13
22	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12
24	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11
27	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10
30	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91
33	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82
36	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75
39	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68
43	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62
47	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56
51	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51
55	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47
62	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43
68	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39
75	36	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36
82	33	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33
91	30	24	20	16	13	11	91	75	62	51	43	36	30	24	20	16	13	11	91	75	62	51	43	36	30
100	27	22	18	15	12	10	82	68	56	47	39	33	27	22	18	15	12	10	82	68	56	47	39	33	27

X is L or C, and tables showing the preferred values for these quantities have already been published[2].

As an example of the use of the tables, let us consider a band-pass filter of 620 ohms impedance and cut-off frequencies 7 000 c/s and 12 000 c/s.

The band-width is 5 000 c/s, giving as the nearest preferred value 5 100 c/s. For this we have already seen that

$$L_{1k} = 39 \text{ mH}.$$
$$C_{2k} = 0.1\mu\text{F}.$$

From Table I we read off the intersection of the preferred cut-off frequencies 6 800 c/s and 12 000 c/s as 30, but as the cut-off frequencies are in different decades we must go to the supplementary table, obtaining $f_m = 9\,100$ c/s.

Looking down the 0.1-μF column of Table II we find that 9 100 c/s corresponds to an inductance $L_{2k} = 3.3$ mH. Looking along the 33-mH row we find that 9 100 c/s corresponds to a capacitance of 0.01 μF, while 100 mH gives a capacitance of 0.0033 μF. Turning to Table III, we obtain the significant figure 82, so that $C_{1k} = 0.0082$ μF. The physical forms for a single section are shown in Fig. 2.

REFERENCES

[1] "Preferred Numbers and Filter Design," *Wireless Engineer*, October 1945.

[2] "Filter Design Tables Based on Preferred Numbers—Low-Pass Filters," *Wireless Engineer*, January 1946.

[3] "Filter Design Tables Based on Preferred Numbers—High-Pass Filters," *Wireless Engineer*, July 1946.

BOOK REVIEWS

Radar Engineering.

By DONALD G. FINK. Pp. 644, with 471 diagrams. McGraw-Hill Publishing Co., Ltd., Aldwych House, London, W.C.2. Price 35s.

Intensive wartime research resulted in vast stores of information on radar, which became very unevenly distributed because of security needs. Mr. Fink's object has been to provide in one volume a review of the whole subject. In doing so he has wisely given prime attention to basic principles, rather than to filling up the book with the endless details of particular systems, which would only have confused the reader approaching the subject for the first time, and obscured the essentials. The author has, in fact, given considerable details of only four sets of equipment, selected to illustrate techniques in the 200, 600, 3,000, and 10,000 Mc/s bands. They are all U.S. Army ground systems, which is perhaps slightly unfortunate in that present-day development is concerned primarily with marine applications. These are hardly mentioned in the book, nor, incidentally, is the "cheese" aerial, which is probably the most important type to-day. Nevertheless, the examples do between them cover a very wide range of technique.

Although the term "radar" is used by the author in its strictest sense, excluding similar systems which do not depend on involuntary echoes, his treatment of the many branches of radio science brought into prominence by radar—such as pulse generation and transmission, waveguides and resonant cavities, and the beaming and propagation of decimetre and centimetre waves—makes much of his teaching applicable not only to pseudo-radar but also to modern developments in television and communications.

The subject-matter is logically and clearly arranged, and concisely expressed, and an unusually full index is provided. References are few because few were available at the time of writing, and the book is by plan self-sufficient. Some of the diagrams (e.g., Figs. 316 and 323) are on too small a scale for comfort; the former, a half-page circuit diagram, includes ten more-or-less complicated waveform pictures, each 0.17 sq cm in area. But with few exceptions the many illustrations are good.

An otherwise commendable work is marred by its elementary sections on circuit principles, notably Secs. 34–37. These purport to explain the current waveforms resulting from voltage pulses, and vice versa; the pulses being derived from generators having respectively zero and infinite internal impedance, applied to circuits containing pure L, C and R, separately and in combination. Many of these cases would not in any event be particularly helpful to the student of radar. For example, a voltage pulse is applied to an inductive circuit having zero resistance. During the pulse the current is quite understandably shown as growing at a uniform rate, E/L. At the end of the pulse the generator is effectively a dead short, with no e.m.f. That being so, it is rather surprising to find that the current is decreasing at the same rate as it grew, apparently without giving rise to any e.m.f. of self-induction; and when it reaches zero it knows it is home and the uniform decay abruptly stops.

With L and R in series, the voltage across the two is shown as dying away exponentially, although the current through them is constant; this phenomenon is matched by a decaying current in parallel C and R with constant terminal voltage. The climax is reached with an incredibly involved description of what is said to happen when a voltage pulse is applied to L and R in parallel. These influences unfortunately crop up elsewhere, as in Fig. 216, where the voltage-current relations in a sawtooth generator appear to be at variance with $e = Ldi/dt$.

Happily this is not the general standard, though no doubt the people who operated the main radar defence system of the British Isles from 1938 to 1945 will deplore the unqualified assertion that beaming is essential to radar. And in the next edition it might be as well to guard against a reasonable impression (from p. 299) that the minimum range of a radar corresponds to pulse duration plus at least several microseconds, for in current practice the determining period is commonly as brief as 0.3 μsec inclusive of pulse. It might also be of interest to trace the career of Mr. Eccles-Jordan! ("The direct-coupled multivibrator . . . is often referred to by his name," p. 330). M. G. S.

The Metre-Kilogram-Second System of Electrical Units.

By R. K. SAS and F. B. PIDDUCK. Pp. 60. Methuen & Co., Ltd., 36, Essex Street, London, W.C.2. Price 4s.

This is one of a series of monographs on physical subjects "intended to supply readers of average scientific attainment with a compact statement of the modern position in each subject." It is a pity that no indication is given of the qualifications of the authors to write such a book; Pidduck we know, but who is Sas? The subject is an important one because, as is mentioned in the introduction, many books on physics in the United States are now written in the m.k.s. system. The authors say that three lines of thought are involved in the discussion, viz., the inconvenience of the present three systems of units, the question whether B and H are of the same nature, and the question of rationalization. The title of the new system is somewhat misleading, for if it were merely a matter of using metres and kilogrammes instead of centimetres and grammes, the only change in any unit would be a power of ten. It is not merely a matter of replacing c.g.s. by m.k.s., but of replacing both c.g.s. (electrostatic) and c.g.s. (electromagnetic) by m.k.s. (what?)

The authors are physicists rather than electrical engineers and an indication of the world in which they move is given by their statement that the microfarad has long been in use, and "now more often the millimicrofarad $m\mu F = 10^{-9}$ F and the picofarad $pF = 10^{-12}$ F." Section 5 of the book is headed "Tribulations of the Student" and we think it peculiarly appropriate. We read "dimensions in the electrostatic system are $m_1^{\frac{1}{2}}l^{\frac{3}{2}}t^{-1}$ and in the electromagnetic system $m^{\frac{1}{2}}l^{\frac{1}{2}}$." As charge had been mentioned several lines back it is presumably to charge that these dimensions refer, but the student will be mystified by the change from m_1 to $m^{\frac{1}{2}}$. Further on in this section it is stated that "the electrical intensity just outside a conductor is $4\pi\sigma S$" which, if S, as we suspect, signifies surface

area, is certainly calculated to add to the tribulations of the student. What Section 7 on "Pulse and Aperture" has to do with the matter it is difficult to see. We are told that "electricians use f as little and ω as much as possible", a statement which we very much doubt, and that "the frequency is measured in cycles"; it is measured in cycles per second, even if the "per second" is often omitted in stating the frequency. They then make the almost incredible proposal to call the unit of ω a pulse (p) so that "50 c = 314 p." They are obviously quite unaware that the term pulsatance is obsolescent and has been banished from the British Standards Glossary. An alternating current of 50 cycles per second may perhaps be regarded as having 50 or 100 pulses per second; it certainly has not 314. Surely *angular frequency* is preferable in every way. In this same section they denote "the divergence of $\mathbf{u}$ by $\nabla\,\mathbf{u}$ and the curl of $\mathbf{u}$ by $\triangle \times \mathbf{u}$." This change from ∇ to $\triangle$ is presumably a printer's error.

It is in Section 8 on Magnetostatics that the tribulations of the student will really begin. Instead of a unit pole being one that repels a similar pole 1 cm away with a force of 1 dyne, it is now one that repels a similar one 1 metre away *in vacuo* with a force of $1/4\pi\mu_0$ newtons where $\mu_0 = 4\pi/10^7 = 1.257 \times 10^{-6}$. The first question that will occur to any reader is whether this is the same unit pole differently defined or a new unit pole, and the reader is left to find out for himself. As a matter of fact, the strength of this new unit pole is about 8 million times that of the classical unit pole. The authors then perpetrate an inexcusable blunder for they take the ampere-turn per metre as the unit of magnetic intensity and call it an *oersted*. One cannot take a name that has been given by international agreement to a certain unit and give it to another unit. One can call a new unit of force a *newton* but certainly not a *dyne*. The statement that "the earth's horizontal component in England is about 15 Oe" is quite inexcusable. This section concludes with the statement that "since flex (*sic*) and pole strength are equal in the m.k.s. system, we can measure pole strength in Wb/m^2." Wb stands for webers but webers per square metre is a measure of B and not of pole strength. There are several more sections dealing with resistance, electromagnetic induction, electrons and the quantum theory, and the book concludes with a list of formulae. In this list we notice "Tension in electromagnetic induction $U = -d\Phi/dt$ where Φ is the total magnetic flux." This strange use of the word tension occurs also in Section 12. We also have "Frequency of electric oscillations $\omega = 1/(LC)^{\frac{1}{2}}$ p," where p stands for pulse.

On p. 53 it is stated that "the unit of magnetic intensity is the *oersted* or $4\pi\mu_0$ times the magnetic intensity at a point 1 m from a unit pole" where μ_0 is the space permeability. Now it is difficult to see how the ratio between two different values of magnetic intensity can be anything but a mere number; it cannot involve the permeability of space. Correctly stated, the oersted, as misused by the authors, is simply $16\pi^2/10^7$ times the value of H 1 metre away from the m.k.s. unit pole *in vacuo* and not, be it specially noted, in a medium of unit permeability. In this rationalized m.k.s. system the permeability of a vacuum is $4\pi/10^7$.

Any student who is still an undergraduate and whose education has been based on the classical systems of units would be very ill advised to look inside this book, for it could lead to nothing but confusion. At a later stage he can please himself.

G. W. O. H.

Second Year Radio Technology

By W. H. DATE, B.Sc., A.M.I.E.E. Pp. 222 x ix. 155 diagrams. Published by Longmans, Green & Co., Paternoster Row, London, E.C.4. Price 7s. 6d.

The author presupposes a knowledge of the basic principles of electric and magnetic circuits and covers elementary radio technology on this basis. Starting with capacitance, inductance and resonant circuits, he goes on to discuss radiation and valves in elementary terms. The use of valves as detectors, amplifiers and oscillators is then dealt with and the book concludes with chapters on direction finding, the superheterodyne and r.f. measuring instruments.

In any elementary book of this nature, the greatest difficulty confronting the author is to achieve accuracy and clarity. If extreme accuracy is aimed at statements are apt to be so hedged about by qualifying clauses that clarity suffers and, if clarity is made the target, there is a danger that the reader takes as of general application statements that are really only true in particular cases. The author succeeds fairly well in avoiding both extremes but in some cases there is danger of giving his readers false impressions.

When treating coupled resonant circuits he omits all reference to circuit resistance and he derives a formula for bandwidth based on the peak separation of resistanceless circuits. There is no indication that the peak separation is affected by resistance nor that, in practice, bandwidth is not usually measured by peak separation but by the difference of frequency between the two points on the resonance curve at which the response bears some definite relation—usually 3 db—to the response at the midband frequency.

In dealing with aerial coupling the author is guilty of several mis-statements of fact. With an inductively loaded aerial he says "If the circuit has a high Q factor the impedance offered to neighbouring unwanted frequencies will be high." In fact, the impedance away from resonance is nearly independent of Q, but the resonance impedance is lower the higher the Q. He should have said that by increasing the current at the resonance frequency the current at neighbouring frequencies is relatively reduced.

With a parallel tuning capacitor he says that the voltage across the coil is Q times the e.m.f. induced in the aerial. It is not; it is $\frac{C_a}{C+C_a}\,Q$ times where C_a and C are the aerial and tuning capacitances respectively.

When treating the circuit in which the aerial is connected to a coil coupled to the tuned circuit proper—the most widely used one in present-day receivers—the author makes some very surprising statements. He says "The effect of the capacitance of the aerial on the tuning circuit is obviated . . ."; "The resistance of the tuning circuit now consists of that of the coil L." He ignores completely the fact that the resistance and reactance of the aerial circuit are reflected into the tuned secondary in an amount dependent on the degree of coupling and that they affect the Q and the tuning of the circuit.

W. T. C.

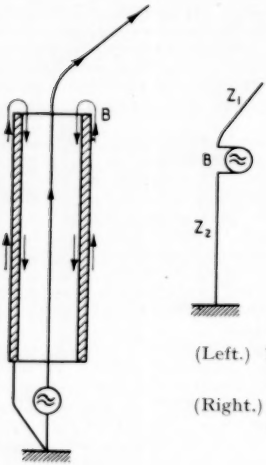
CORRESPONDENCE

Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Partially-Screened Open Aerials

To the Editor "Wireless Engineer."

SIR,—I beg leave to query the fundamental concepts of the article on this subject by Mr. Burgess which appeared in your number of May last.



(Left.) *The two faces of the screening.*

(Right.) *Symbolic form of actual aerial.*

I believe it is essential to consider the two faces of the screening separately (always provided that the thickness of the screening is much greater than the depth of penetration corresponding to the working frequency) since the internal face forms, together with the central conductor, a balanced co-axial feeder; while the external face together with the prolongation of the aerial acts as the actual aerial, being fed from the co-axial feeder at the point B.

It is therefore evident that the screening also acts as a collector of the electric field if the aerial is used for reception.

If this exception be admitted, the calculations may be carried out as stated by Mr. Burgess in his article, and a first approximation may be obtained by comparing the various lengths to lines of differing impedance.

A. COLINO.

Tech. Director, Marconi Española, S.A.
Madrid.

Doppler Effect in Propagation.

To the Editor, "Wireless Engineer."

SIR,—In the article on the "Doppler Effect in Propagation," by Mr. H. V. Griffiths, in the June 1947, issue of *Wireless Engineer*, emphasis is laid on the difference between the classical values of the frequency shift according to whether the source is moving towards the observer or the observer towards the source. If, however, this difference were real it would be possible to distinguish states of absolute rest or motion by the measurement of the frequency shift. However the relativity principle states that this is impossible and that, quite generally, observations made in two frames of reference in uniform relative motion can only determine their *relative* velocity.

From the equations of special relativity giving the transformation from one frame of reference to another it is readily shown that the ratio of the observed frequency f_0 to the source frequency f_s when they are in uniform relative motion towards each other with velocity v is given by

$$\frac{f_0}{f_s} = \sqrt{\frac{c+v}{c-v}}$$

(See, e.g., J. H. Jeans "Electricity and Magnetism," 5th Edition, p. 609.) There is thus no question of motion of the source and observer relative to the "ether" but only of their motion relative to each other. In the case of sound waves the problem is different since the medium propagating the waves is in this case material and therefore the terms "rest" and "motion" can be given an absolute meaning.

The difference between the classical values of the frequency ratio (given by the author's equations 1 and 2) and the correct value given above is negligible in the radio problem discussed. However, it seemed worth while to raise the matter especially as several well-known text books on optics still give the classical values and discuss their difference as though it were real.

At the end of the article it is stated that $\Delta H/\Delta l$ is difficult to calculate whereas in fact from simple geometrical considerations it is seen to be equal to sec β which for the numerical examples given is equal to 3.5 in the two-hop case and 3.0 in the three-hop case; the values given by the author are rather low.

R. E. BURGESS.

National Physical Laboratory,
Teddington, Middx.

To the Editor, "Wireless Engineer."

SIR,—I was astonished to read in the article by H. V. Griffiths in the June issue that the Doppler effect gives observed frequencies which depend upon the *absolute* velocity of the source and observer, not just upon the relative velocity between them, a statement upon which the author laid particular stress. This statement recalls the age of classical physics when many experimenters tried to measure the absolute velocity of the earth in "ether." All experiments, notably those of Michelson and Morley, have proved the principle of the special theory of relativity that absolute motion has no meaning.

According to Griffiths, if the earth E were moving with a velocity v_E, a light-emitting star S with v_S (both v_E and v_S approaching along the line ES), then the frequency observed at the earth is

$$f = f_S \frac{c + v_E}{c - v_S}$$

Let $k = f_S/f$ = source freq./observed freq., then $k(c + v_E) = c - v_S$, and $v_S + kv_E = c(1 - k)$. Relative velocity of the star as observed from the earth $= v_E + v_S = (1 - k)(c + v_E)$.
(Applying the classical addition of velocities). This depends upon the absolute velocity of the earth in "ether."

The difficulty may be easily overcome by the application of the special theory of relativity, in particular the Lorentz transformation between coordinate systems moving relatively at a uniform velocity.

Let us have two coordinate systems, one C is fixed to the earth, and the other C' is fixed to the star. Let C move relative to C' at a velocity v in the X direction. The Z axis is taken perpendicular to the plane defined by the X axis and the line ES. The symbols f and f_s will be changed into f' and f to suit the new notation. The line ES makes an angle α with the X direction.

The phase of the light in system C is

$$\left(e^{2\pi jf\left(t-\frac{X\cos\alpha+Y\sin\alpha}{c}\right)}\right).$$

That in C' is $\left(e^{2\pi jf'\left(t'-\frac{X'\cos\alpha+Y'\sin\alpha'}{c}\right)}\right)$. The phase should be invariant in both systems, hence

$$f\left(t-\frac{X\cos\alpha+Y\sin\alpha}{c}\right) = f'\left(t'-\frac{X'\cos\alpha'+Y'\sin\alpha'}{c}\right).$$

On applying the laws of transformation of X, Y, Z, t into X', Y', Z', t', we get

$$f'=f\frac{1-\beta\cos\alpha}{\sqrt{1-\beta^2}} \qquad \text{where } \beta=\frac{v}{c}$$

$$\text{and } \tan\alpha'=\frac{\sin\alpha\sqrt{1-\beta^2}}{\cos\alpha-\beta}$$

When $\alpha=0$ then $f'=f\sqrt{\frac{1-\beta}{1+\beta}}$ and $\alpha'=0$

If $\beta \ll 1$ then $f'=f(1-\beta)$, an approximation which coincides with both formulae given by Griffiths for speeds small compared with c, independent of any absolute motion. If $v=c$, $\beta=1$ and $f/f'=$ observed freq./source freq. $=\infty$, independent of any absolute motion, not once ∞ and once 2 according to which system is the one fixed in space.

If $\alpha=90°$ (relative motion perpendicular to ES) then $f'=f\frac{1}{\sqrt{1-\beta^2}}$, and $\tan\alpha'=-\sqrt{\frac{1}{\beta^2}-1}$.

Of course the relativity calculation of the Doppler effect was well known long ago.[1] I write this just to draw attention to the fact.

F. S. ATIYA.

Federal Institute of Technology,
Zürich.

[1] W. Pauli, Jun., "Relativitätstheorie."

To the Editor, "Wireless Engineer."

SIR,—The article by H. V. Griffiths[1] on the Doppler Effect on Propagation is of interest from two points of view, the application of the effect to a study of the movements of the reflecting layers, and its influence on the accuracy of comparing the frequencies of distant standards. From both points of view a complete record of frequency throughout the period of observation would be more useful than the average value. When this manifestation of the Doppler Effect was first noticed[2] it was found that the frequency of the received signal was steady for long periods but was subject to large variations for relatively short periods. From the nature of the observations it was decided that for the purpose of comparing the distant standards it was better to ignore the large variations, which could only be caused by rapid fluctuations in the conditions of propagation, and it was estimated that the result of the comparison was then accurate to $\pm$ 2 parts in 10^8. This could not be checked at the time because the frequency of the standard used was not known with the necessary accuracy, but was confirmed by a series of comparisons carried out in 1938. Mr. Griffiths does not give the results of his recent measurements in detail and it is therefore not possible to say whether some of the deviations could justifiably be neglected for the purpose of frequency comparison. It would be interesting to know if appreciable errors arise when the received signal gives steady uniform beats with the local standard, that is, if the reflecting layers move uniformly for long periods of time, or whether large movements are always associated with unsteady conditions of reception and a variable frequency as my earlier experiments suggested.

L. ESSEN.

National Physical Laboratory,
Teddington, Middx.

[1] H. V. Griffiths. "Doppler Effect in Propagation." *Wireless Engineer*, Vol. XXIV, p. 162, June 1947.

[2] L. Essen. "International Frequency Comparisons by Means of Standard Radio Frequency Emissions." *Proc. Roy. Soc.*, April, 1935, A.149, p. 506.

Permeability of Dust Cores

To the Editor, "Wireless Engineer."

SIR,—One or two remarks in Dr. Friedlaender's letter in your June issue seem to call for a reply and I would offer the following comments :—

(1) It was not explicitly stated in my letter (February, 1947) that slabs of dimension ratio, 10 : 2 : 1 had been observed, but that permeability curves based on these dimensions gave an approximately true picture of the behaviour of the higher permeability cores. Some particles that have been examined do approach these relative dimensions, but it will of course be appreciated that some of the flattening in such cases might have occurred prior to pressing. In most cases the ratio is smaller.

(2) There are many factors, including those mentioned by Dr. Friedlaender, which contribute to the value of effective permeability of a dust core. Particles in alloy cores are often of irregular outline and a rigid mathematical treatment is hardly practicable. The case in which some of the insulation becomes trapped without seriously impeding the flux was catered for to some extent in my curve (2) for truncated cubic particles.

(3) My permeability figures for alloy powder cores, pressed with no insulation are almost certainly of the right order since they were obtained with commercial powders and separate cross-checks on the a.c. and d.c. methods of measurement were carried out. The fact that the permeability of such cores is not the same as the permeability of the metal was implied in my quoting the two values of 1000 and 3000 for cores with considerable air-spaces and with few air-spaces respectively.

P. R. BARDELL.

G.E.C. Research Laboratories,
Wembley.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each.

AERIALS AND AERIAL SYSTEMS

582 328.—Waveguide with a slotted end-wall which serves as a dipole to radiate a beam of energy.

The British Thomson-Houston Co. (communicated by the General Electric Co.). Application date 2nd November, 1943.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

581 201.—Frequency-indicating device, particularly for radiolocation, wherein the signal is passed through two electrical paths which introduce a frequency, phase, or amplitude distortion the sum and difference of which are applied in quadrature to a c.r. oscilloscope.

Standard Telephones and Cables Ltd. and C. W. Earp. Application date 21st April, 1942.

581 561.—Pulse-generating circuit, with means for rapidly damping the oscillations to facilitate close-range observations in radiolocation.

A. D. Blumlein and E. L. C. White. Application dates 17th June, 1940, *and 23rd January,* 1941.

581 568.—Radiolocation equipment of the "triggered" or i.f.f. type in which a variable delay is introduced before re-transmission, to baffle unauthorized receivers.

Standard Telephones and Cables Ltd. and B. B. Jacobsen. Application date 14th November, 1941.

581 576.—Rotating compass beacon in which two overlapping beams form an equi-signal line, along which a masking signal becomes zero and a bearing signal is heard.

W. A. S. Butement and L. H. Bedford. Application date 28th January, 1944.

581 602.—Navigational system in which a given course is steered by maintaining a constant time-interval between the pulsed signals received from two or more spaced transmitters.

R. J. Dippy. Application date 23rd December, 1942.

581 603.—Control and synchronization of master and satellite stations radiating pulsed signals which form a navigational lattice in space.

R. J. Dippy. Application date 23rd December, 1942.

581 920.—Radiolocation system utilizing a modulated carrier-wave and two or more spaced pick-up devices feeding a cathode-ray indicator.

A. D. Blumlein and L. C. White. Application date 20th July and 3rd November, 1939.

581 980.—Radiolocation equipment to facilitate the rapid alignment of a searchlight projector on a moving target.

J. D. Cockcroft, D. R. Chick, W. S. Eastwood and A. J. H. Oxford. Application date 11th December, 1942.

581 982.—Radiolocation apparatus in which a super-regenerative oscillator serves both as a transmitter and receiver, the quenching-frequency indicating the echo-time of the exploring signal.

A. C. Cossor, Ltd. and F. R. W. Strafford. Application date 30th December, 1942.

581 989.—Radiolocation apparatus in which a "gate" switch operated at the pulsing frequency serves to prevent jamming of the set by c.w. waves.

J. D. Cockcroft and A. J. H. Oxford. Application date 17th September, 1943.

581 996.—Radiolocation apparatus in which a periodic control-voltage allows the echo-signals from a selected target to be distinguished from other echo-signals.

J. D. Cockcroft, D. R. Chick, W. S. Eastwood and A. J. H. Oxford. Application date 11th December, 1942.

581 998.—Radiolocation aerial-system comprising four units grouped around a common axis and a common coupling-ring for constantly rotating the lobe of maximum directivity.

J. D. Cockcroft, D. R. Chick, W. S. Eastwood and A. J. H. Oxford. Application date 11th December, 1942.

582 171.—Radiolocation equipment for indicating to each of a pair of pursuit craft which of the two is nearer to a common target or pursued craft.

Standard Telephones and Cables, Ltd., P. K. Chatterjea and L. W. Houghton. Application date 17th July, 1942.

582 325.—Radiolocation system in which the echo-signals are received by three different aerial systems to give coordinated indications of the distance, bearing, and elevation of the reflecting body or target.

J. D. Cockcroft and P. E. Pollard. Application date 24th March, 1942.

582 360.—Impulse-generator for controlling the fly-back stroke of the cathode-ray indicator in radiolocation equipment.

Marconi's W.T. Co., Ltd., R. J. Kemp and D. J. Fewings. Application date 28th May, 1940.

582 403.—Radiolocation indicator in which the time-base makes several sweeps between successive pulses in order to provide an open zigzag scale.

Marconi's W.T. Co., Ltd., R. J. Kemp and D. J. Fewings. Application date 28th May, 1940.

582 419.—Radiolocation equipment, particularly for following the movement of surface craft at sea, the range being shown against a spiral time-base on one c.r. tube, and the bearing on a linear time-base on a separate tube.

J. D. Cockcroft, J. Ashhead, E. Coop, A. E. Kempton, P. G. Forsyth and B. Newsam. Application date 30th June, 1943.

582 434.—Generating pulses, say for radiolocation, by means of a self-blocking oscillator valve, and an associated column of liquid of variable length, backed by a pair of piezoelectric crystals.

Scophony, Ltd. and A. F. H. Thomson. Application date 15th May, 1944.

582 466.—Overlapping-beam approach-path and blind-landing system, in which provision is also made for a one-way telephone channel between ground and air.

Standard Telephones and Cables, Ltd. and H. P. Williams. Application date 16th May, 1944.

582 493.—Radiolocation equipment, particularly for detecting ships and low-flying aircraft, in which the direction of exploration is constantly varied over a given angle.

J. D. Cockcroft, J. Ashmead, W. S. Eastwood, A. J. H. Oxford, A. F. H. Thomson and W. A. S. Butement. Application date 4th December, 1942.

582 503.—Radiolocation apparatus in which a variable-range strobing voltage is applied to search-out and isolate a selected echo-signal from other interfering signals: applicable to aircraft-interception.

F. C. Williams, E. L. C. White and D. Blumlein (legal representative of A. D. Blumlein). Application date 15th October, 1943.

582 708.—Radiolocation system of the variable-frequency type in which the beat-frequency, normally produced is converted into a pure sine wave.

Standard Telephones and Cables, Ltd. and C. W. Earp. Application date 3rd April, 1942.

582 755.—Radio beacon comprising a pair of non-directional aerials, mounted on a common rotating base, and energized through phase-shifting and switching circuits.

Soc. Anon des Industries Radio-Electriques S.A.D.I.R. Convention date (France), 29th May, 1941.

582 768.—Mechanical and electronic switching devices for a waveguide feeding, say, several aerials in succession, as in radiolocation.

Marconi W.T. Co., Ltd. (assignees of W. D. Hershberger). Convention date (U.S.A.) 25th February, 1943.

582 850.—Pulse-generating circuit, say for radio-location, in which a modulating voltage is applied by discharging a capacitance storage network through triggered spark-gaps.

The British Thomson-Houston Co., Ltd. (communicated by the General Electric Co.). Application date, 8th November, 1943.

RECEIVING CIRCUITS AND APPARATUS

581 913.—Super-regenerative circuit, with a variable quenching-control, particularly for receiving television or pulse-modulated signals.

The British Thomson-Houston Co., Ltd. Convention date (U.S.A.) 3rd February, 1943.

581 976.—Mixing circuit for a superheterodyne receiver in which a crystal rectifier is coupled to a waveguide through coaxial-line elements.

E. C. Cork, M. Bowman-Manifold, F. H. Gale and R. Blythen. Application date 1st September, 1942.

582 127.—Tuning arrangement for a receiver of the "panoramic" type where the whole of a selected wave-band is scanned, the signals being recorded against the time-base of a c.r. tube.

H. Whalley, C. W. Miller and Metropolitan-Vickers Electrical Co., Ltd. Application date 8th September, 1941.

582 128.—Switch-tuning for a multi-waveband receiver of the "panoramic" type.

J. M. Dodds and Metropolitan-Vickers Electrical Co., Ltd. Application date 8th September, 1941.

582 131.—Tuning arrangement for a multi-waveband receiver of the "panoramic" type.

G. J. Scoles, H. Whalley and Metropolitan-Vickers Electrical Co. Ltd. Application date 8th November, 1941.

582 617.—Means for reducing "shot noise" in an electronic resonator, used say as a velocity-modulation mixer in a superheterodyne receiver.

W. S. Percival. Application date 25th July, 1941.

TRANSMITTING CIRCUITS AND APPARATUS

581 984.—Concentric line impedance transformer in which the effective length of a right-angled stub is controlled by a sliding piston.

Sperry Gyroscope Inc. (assignees of W. W. Hansen and J. R. Woodyard). Convention date (U.S.A.) 17th May, 1941.

582 088.—Means for neutralizing the undesired secondary waves that are generated at a bend or angle in a wave-guide.

E. C. Cork and M. Bowman-Manifold. Application date 14th August, 1942.

582 585.—Automatic tuning-control for a radio-transmitter wherein a tuning reactance is regulated, through a motor, by the phase-difference between input and output voltages.

Marconi's W. T. Co. Ltd. (assignees of O. B. Cunningham, H. Sussman and S. Gubin). Convention date (U.S.A.) 29th March, 1943.

582 757.—The use of localized plates and like impedance-varying elements to prevent undesired reflection-effects when energy flows through a waveguide containing sections with different dielectric-constants.

The British Thomson-Houston Co. Ltd. Convention date (U.S.A.) 16th April, 1942.

582 766.—Intercommunication set comprising a super-regenerative detector which serves (a) to receive frequency or amplitude modulated signals and (b) to generate a carrier-wave for f.m. outgoing signals.

Marconi's W. T. Co. Ltd. (assignees of M. G. Crosby). Convention date (U.S.A.) 15th February, 1943.

582 795.—Transmitter circuit in which the anode-cathode paths of two valves are in series resonance with a crystal driver, one of the valves serving as an amplifier, modulator, and start-stop switch.

D. L. Hings. Application date 23rd July, 1942.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

582 100.—System for translating amplitude-modulated signals into equivalent time-modulated pulses.

Standard Telephones and Cables Ltd. (assignees of D. D. Greig). Convention date (U.S.A.) 30th April, 1943.

582 146.—Triggered pentode oscillator-circuit for generating trains of pulses of constant amplitude throughout each repetition-period.

Standard Telephones and Cables Ltd. (assignees of L. A. de Rosa). Convention date (U.S.A.) 30th November, 1942.

582 709.—System of pulsed signalling wherein secrecy is secured by the use of an amplitude-modulated masking-wave, which is removed at the receiving end (addition to 580 843).

Standard Telephones and Cables Ltd., P. K. Chatterjea and L. W. Houghton. Application date 23rd October, 1942.

582 798.—Secret signalling system in which a carrier-wave is frequency-modulated both by the desired signal and by a distorting wave, the latter being removed at the receiving end.

The British Thomson-Houston Co. Ltd. (communicated by the General Electric Co.). Application date 25th November, 1943.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

581 895.—Reflecting electrode for a rhumbatron resonator adapted to accommodate a wide variation of the biasing potential.

A. F. Pearce. Application date 16th December, 1941.

582 021.—Reducing interelectrode capacitance in short-wave amplifiers of the kind in which the grid and anode leads consist of disks extending through the glass bulb.

Standard Telephones and Cables Ltd. (assignees of P. G. Chevigny). Convention date (U.S.A.) 24th December, 1942.

582 162.—Cathode construction designed to reduce the electron transit-time in a short-wave valve.

The M-O Valve Co. Ltd. and G. W. Warren. Application date 13th March, 1940.

582 171.—Oscillation-generator of the velocity-modulation type comprising a variable-mu control grid to facilitate tuning adjustment.

Standard Telephones and Cables Ltd. and R. G. Roach. Application date 8th May, 1942.

582 222.—Short-wave oscillator valve in which the coupled circuits consist of telescopic transmission-line elements mounted on a common axis.

The British Thomson-Houston Co. Ltd. (communicated by the General Electric Co.). Application date 25th May, 1944.

582 318.—Short-wave oscillator-unit comprising a valve mounted inside a pair of inter-coupled transmission-line elements.

Standard Telephones and Cables Ltd. (communicated by International Standard Electric Corporation). Application date 21st September, 1944.

582 484.—Velocity-modulation tube in which the rims of the resonator-elements are "chiselled" to minimize the effect of secondary emission.

The British Thomason-Houston Co. Ltd. and C. J. Milner. Application date 29th October, 1941.

582 487.—Velocity-modulation devices in which the tuning of the resonator elements is controlled by a differential screw mechanism.

The British Thomson-Houston Co. Ltd. and W. J. Scott. Application date 22nd December, 1941.

582 618.—Construction and arrangement of the filament and grid wires, to minimize "hum" in a valve intended to be driven from an a.c. source.

Standard Telephones and Cables Ltd. and A. J. Maddock. Application date 26th July, 1941.

582 935.—Construction and arrangement of the resonant cavities in an oscillation-generator utilizing velocity-modulation.

Standard Telephones and Cables Ltd., J. H. Fremlin and J. Foster. Application date 19th September, 1941.

582 936.—Resonant-cavity device, particularly for generating ultra-high frequencies, consisting of a stack of apertured plates (divided out of 582 935).

Standard Telephones and Cables Ltd., J. H. Fremlin and J. Foster. Application date 19th September, 1941.

583 024.—Preventing electronic damping effects in velocity-modulation and other electron-beam devices.

Standard Telephones and Cables Ltd. and S. G. Tomlin. Application date 2nd May, 1941.

583 041.—Construction and sealing of electron discharge devices and spark-gaps, where the envelope is made of ceramic material.

The British Thomson-Houston Co. Ltd. and H. de B. Knight. Application date 10th November, 1943.

583 180.—Screw-type tuning adjustment for the cavity-resonator of an electron-discharge device of the velocity-modulation type.

Standard Telephones and Cables Ltd., J. H. Fremlin and S. G. Tomlin. Application date 6th December, 1943.

583 213.—Construction and spacing of the electrode system of a Klystron oscillation-generator, designed to combine easy starting with high operating efficiency.

Standard Telephones and Cables Ltd. and J. H. Fremlin. Application date 12th December, 1941.

583 214.—Electrode construction and arrangement for a discharge device of the velocity-modulating type, particularly where only one cavity-resonator is used.

Standard Telephones and Cables Ltd., J. H. Fremlin and C. Strachey. Application date 12th December, 1941.

584 051.—Construction and arrangement of wire-gauze radiating members forming part of the cooling system of a transmitting valve.

Standard Telephones and Cables Ltd. and A. S. Wade. Application date 23rd November, 1944.

SUBSIDIARY APPARATUS AND MATERIALS

582 018.—Phase-reversing amplifier for obtaining a variable output proportional to the difference between two variable potentials, applicable to calculating-apparatus.

A. C. Cossor Ltd. and L. Jofeh. Application date 12th May, 1944.

583 246.—Frequency-changer stages, with coupling networks, for analyzing a complex wave or for calibrating a wide range of frequencies.

Standard Telephones and Cables Ltd. and C. T. Daly. Application date 14th July, 1944.

583 553.—Pentode valve circuit including a regenerative diode switch on the control-grid for developing a square-shaped wave-form at predetermined times.

F. C. Williams. Application date 2nd October, 1944.